elektor

up-to-date electronics for lab and leisure

ASCII keyboard

plus

elekterminal

equals

TV typewriter

elektor 44 decoder

What is a TUN?
What is 10 n?
What is the EPS service?
What is the TQ service?
What is a missing link?

Semiconductor types
Very often, a large number of
equivalent semiconductors exist
with different type numbers. For
this reason, 'abbreviated' type
numbers are used in Elektor
wherever possible:
- '741' stand for μA741,
LM741, MC641, MIC741,
RM741, SN72741, etc.
- 'TUP' or 'TUN' (Transistor,
Universal, PNP or NPN respect-
ively) stand for any low fre-
quency silicon transistor that
meets the following specifi-
cations:

UCEO, max	20V
IC, max	100 mA
hfe, min	100
Ptot, max	100 mW
fT, min	100 MHz

Some 'TUN's are: BC107, BC108
and BC109 families; 2N3856A,
2N3859, 2N3860, 2N3904,
2N3947, 2N4124. Some 'TUP's
are: BC177 and BC178 families;
BC179 family with the possible
exeption of BC159 and BC179;
2N2412, 2N3251, 2N3906,
2N4126, 2N4291.

- 'DUS' or 'DUG' (Diode Univer-
sal, Silicon or Germanium
respectively) stands for any
diode that meets the following
specifications:

	DUS	DUG
UR, max	25V	20V
IF, max	100mA	35mA
IR, max	1µA	100 µA
Ptot, max	250mW	250mW
CD, max	5pF	10pF

Some 'DUS's are: BA127, BA217,
BA218, BA221, BA222, BA317,
BA318, BAX13, BAY61, 1N914,
1N4148.
Some 'DUG's are: OA85, OA91,
OA95, AA116.

- 'BC107B', 'BC237B', 'BC547B'
all refer to the same 'family' of
almost identical better-quality
silicon transistors. In general,
any other member of the same
family can be used instead.

BC107 (-8, -9) families:
BC107 (-8, -9), BC147 (-8, -9),
BC207 (-8, -9), BC237 (-8, -9),
BC317 (-8, -9), BC347 (-8, -9),
BC547 (-8, -9), BC171 (-2, -3),
BC182 (-3, -4), BC382 (-3, -4),
BC437 (-8, -9), BC414

BC177 (-8, -9) families:
BC177 (-8, -9), BC157 (-8, -9),
BC204 (-5, -6), BC307 (-8, -9),
BC320 (-1, -2), BC350 (-1, -2),
BC557 (-8, -9), BC251 (-2, -3),
BC212 (-3, -4), BC512 (-3, -4),
BC261 (-2, -3), BC416.

Resistor and capacitor values
When giving component values,
decimal points and large numbers

of zeros are avoided wherever
possible. The decimal point is
usually replaced by one of the
following abbreviations:

p	(pico-)	=	10^{-12}
n	(nano-)	=	10^{-9}
μ	(micro-)	=	10^{-6}
m	(milli-)	=	10^{-3}
k	(kilo-)	=	10^{3}
M	(mega-)	=	10^{6}
G	(giga-)	=	10^{9}

A few examples:
Resistance value 2k7: 2700 Ω.
Resistance value 470: 470 Ω.
Capacitance value 4p7: 4.7 pF, or
0.000 000 000 004 7 F . . .
Capacitance value 10n: this is the
international way of writing
10,000 pF or .01 μF, since 1 n is
10^{-9} farads or 1000 pF.
Resistors are ¼ Watt 5% carbon
types, unless otherwise specified.
The DC working voltage of
capacitors (other than electro-
lytics) is normally assumed to be
at least 60 V. As a rule of thumb,
a safe value is usually approxi-
mately twice the DC supply
voltage.

Test voltages
The DC test voltages shown are
measured with a 20 kΩ/V instru-
ment, unless otherwise specified.

U, not V
The international letter symbol
'U' for voltage is often used
instead of the ambiguous 'V'.
'V' is normally reserved for 'volts'.
For instance: U_b = 10 V,
not V_b = 10 V.

Mains voltages
No mains (power line) voltages
are listed in Elektor circuits. It is
assumed that our readers know
what voltage is standard in their
part of the world!
Readers in countries that use
60 Hz should note that Elektor
circuits are designed for 50 Hz
operation. This will not normally
be a problem; however, in cases
where the mains frequency is used
for synchronisation some modifi-
cation may be required.

Technical services to readers
- EPS service. Many Elektor
articles include a lay-out for a
printed circuit board. Some — but
not all — of these boards are avail-
able ready-etched and predrilled.
The 'EPS print service list' in the
current issue always gives a com-
plete list of available boards.
- Technical queries. Letters with
technical queries should be
addressed to: Dept. TQ. Please
enclose a stamped, self addressed
envelope; readers outside U.K.
please enclose an IRC instead of
stamps.
- Missing link. Any important
modifications to, additions to,
improvements on or corrections
in Elektor circuits are generally
listed under the heading 'Missing
Link' at the earliest opportunity.
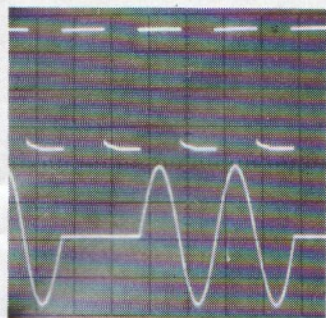
A (repetitive) tone-burst signal is an extremely useful aid for testing audio equipment. Basically, this type of test signal is obtained by switching the output of a sine-wave oscillator on and off at regular intervals. The **tone-burst generator** utilises a novel design approach that simplifies the circuit considerably and only involves one minor reduction of the capabilities.
p. 12-10
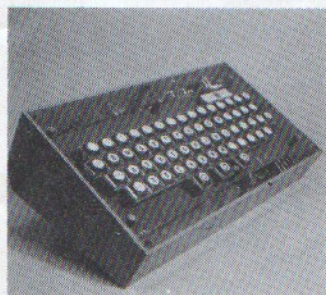
The most convenient and elegant method of displaying data from a micro-processor is on a Visual Display Unit (VDU). When used in conjunction with an ASCII keyboard, a video interface circuit forms a complete video data terminal.
The **elekterminal** can be used with the Elektor SC/MP system or any microprocessor system possessing a serial input/output facility.
p. 12-16
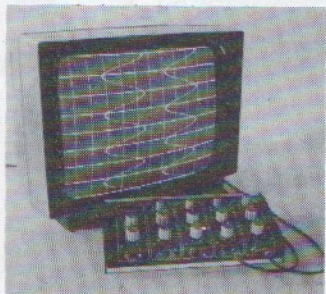
Although the basic version of the **TV scope** is an extremely useful tool when testing low-frequency circuits, the **extended version** offers vastly superior display possibilities. Any signal within the frequency range of the scope (DC to 100 kHz) can be displayed as a clear and stable trace on the screen of any domestic TV receiver.
p. 12-34

This month's cover illustrates the close relationship between the ASCII keyboard, published last month, and the Elekterminal described in this issue. Although they are two independent units, each useful in its own right, the combination is equal to a 'TV typewriter' with a microprocessor-compatible interface.

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SELEKTOR

Major development in telephone engineering

A fivefold increase in the capacity of existing telephone lines might be possible within five years as a result of a new speech processing system being developed in Britain.

The UK Ministry of Defence says a top military communications expert and a team from the University of Bath's School of Electronic Engineering in Western England has solved a speech transmission problem that has baffled scientists for years.

The result, it is claimed, is a system that transmits speech by wire or radio with greater clarity and efficiency than at present. It will be marketed, it is hoped, at less than a tenth of the cost of existing systems.

Patent applications have already been filed and experts are now predicting that worldwide commercial sales could be worth millions of pounds sterling. It is not only the world's telephone systems that stand to gain from the development. Others who could benefit, in the long-term, include deep sea divers and the partially deaf. In the case of divers the system could improve the quality of helium-affected speech, while hearing aids could be made more efficient.

Terminal smaller than shoebox

The Ministry of Defence says details of how the new system works are classified. But it is known that it reduces speech patterns with the aid of computers to a small number of basic 'shapes.' It is possible to break down and reconstruct combinations of these shapes like jigsaw puzzles.

The system's terminal will be smaller than a shoebox.

The military communications expert involved, Brigadier Reg King, did some research into speech processing eight years ago but it was not until he joined forces last year with Professor William Gosling, head of Bath University's School of Electronic Engineering, that major progress was made.

Brigadier King said: 'In the end, the answer was so simple that I had trouble in convincing myself that it really worked. I had the system checked on a second computer in case we had overlooked some vital detail, but it came up with the same result'.

Professor Gosling commented: 'There is no doubt whatsoever that this represents a revolutionary concept in communications engineering. The mathematicians have always said that it could not be done. I think it was mainly Brigadier King's ability to tackle the problem obliquely and his dogged

military persistence in refusing to accept defeat that enabled him to solve the seemingly insoluble'.

A prototype of the system now being built should be ready for evaluation in some 18 months time. It is said that a fully engineered version could be in general use within the next five years.

(361 S)

New Finds

A new smaller-diameter version of the Ferranti inertial directional surveyor (FINDS), known as the Model 1063, has been developed. The outside diameter of the new tool is $10^{5}/_{8}$ inches (270 mm) which will enable it to be used to survey $13^{3}/_{8}$ inches casings down to depths of 8,000 ft.

Preliminary trials have indicated that the FINDS 1063 tool can produce survey measurements to an accuracy of 6 inches in all three channels (North, East and vertical) at the maximum depth at which it has been used to date — 4,800 ft. The trials have been conducted in association with Shell Expro in the Brent and Dunlin fields. The first experimental surveyor designed by Ferranti had a diameter of $17^{1}/_{2}$ inches and was used down to depths of 1,500 ft. very successfully. The significantly smaller diameter of the new 1063 tool has been achieved by repackaging the inertial sensor assembly and its associated electronic units. This will enable oil and gas wells to be surveyed very accurately over a significant proportion of their total depth.

FINDS - principle of operation

FINDS is an adaptation of an inertial

sensing unit designed for the precise navigation of aircraft. The unit has been proven in service over many years and is exceedingly robust.

The inertial sensing unit comprises a gyroscopically stabilised platform which is maintained within one minute of arc of a fixed attitude in space regardless of changes in the orientation of the vehicle in which it is being carried. Three precision accelerometers are mounted on the stabilised platform with their axes aligned mutually at right angles to one another. One axis is automatically gyro-compassed to face north. The accelerometers detect accelerations along all three axes and the output signals are integrated twice to derive displacements from the known starting datum. These functions are carried out whilst the tool is in motion and the derived co-ordinates recorded in a semi-conductor store every 0.1 second.

On recovery of the tool these results are transferred to a computer store for immediate analysis, the results being produced at specified depth intervals and presented in a chosen format. Battery powered, the entire system is self-contained and can operate for several hours independently of external services. Survey runs, however, are of short duration since the tool can be lowered at the maximum wireline (or sandline) speed and has been tested in free-fall conditions at 1,500 ft/min. It need only be stopped for ten second fixes every one to two minutes and this transit interval can be increased, if required, at the expense of accuracy.

Ferranti Offshore Systems Ltd.
Ferry Road
Edinburgh
EH5 2XS

(358 S)
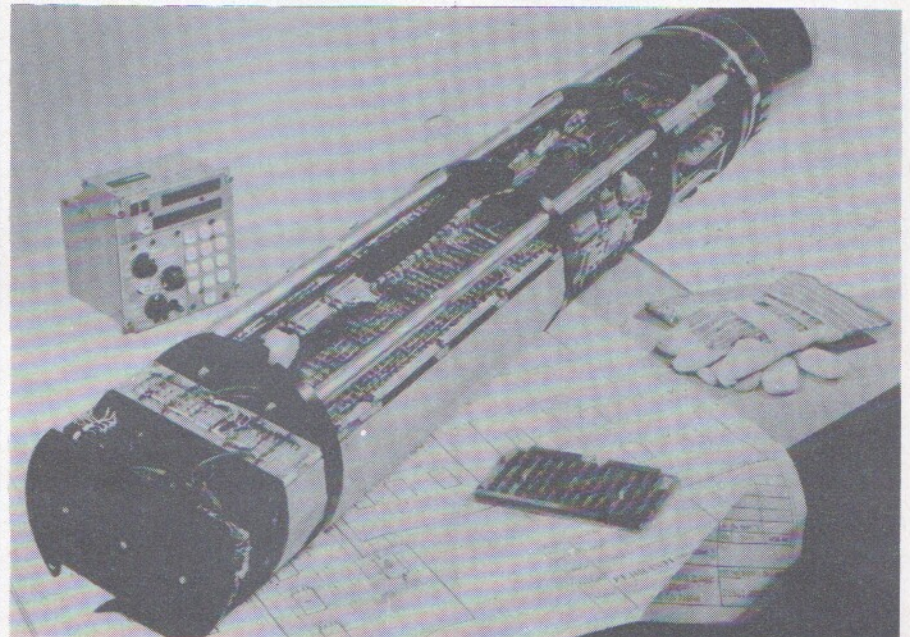
SELEKTOR

Telephone with telescribble

People will soon be able to speak and write to each other via the telephone at the same time. Philips Telecommunicatie Nederland has developed a system, called the Scribofoon, comprising an electronic note pad and a visual display unit, which can be used to transmit graphic information via telephone lines. The system can be used either by two people, by several people if they are taking part in a telephone conference, and in mobilophone communications. A number of pilot projects will be set up in the Netherlands next year (1979) to evaluate the system.
Once the Scribofoon has become widely accepted, its cost is expected to be about the same as that of a colour television set. Before that, however, it will cost more because of the small series which will be made.

Philips Telecommunicatie Nederland based the Scribofoon's development on research done at Delft University of Technology where the staff of the Information Transmission Laboratory developed a method of transmitting pictures via radio. The work done by Philips Telecommunicatie Nederland, in co-operation with the Dutch Post Office's Dr. Neher Laboratory, aimed at applying this method to achieve two-way picture transmission via telephone lines and at developing the equipment. A very small part of the telephone frequency band, the conversation band, is reserved for transmitting the pictures without any unfavourable influence on the telephone conversation. This means that picture and speech can be transmitted simultaneously so that one can talk and write via the telephone at the same time. Thus, a telephone conversation can be clarified with drawings,

sketches and maps for example. When one person is writing, the other cannot do so because the facility is blocked. The Scribofoon consists of a screen and a writing pad. This pad contains printed wiring: vertical wires on top of a plastic layer and horizontal wires underneath it. An electric pulse scans this network of wires. When a pen touches the paper covering the pad it acts as an antenna and signals the moment when the rapidly scanning pulse is maximal, in other words nearest to the pen. The system then knows the pen's position because the time-lag between the start of the pulse and it being seen by the pen is specific for a certain position on the pad. These positions are then transmitted via the telephone line and made visible on the other subscriber's screen. This transmission is made at normal writing speed.
When a group of people are using the system, the graphic information appears on everyone's screen at the same time. Members of the group can propose additions or corrections because an erase facility is a feature of the Scribofoon. The pen can also be used to point out a detail of the picture without this mark being incorporated into the picture. Both the telephone drawing and the spoken conversation can be recorded on a normal cassette recorder for later reproduction. In the Scribofoon's application with mobilophones, bodies such as the police and fire brigade could use the system for quickly transmitting a situation sketch, for example.
International interest has already been shown in the system, also for its possible uses in education. A pre-recorded, language-teaching cassette could be used to pass on spoken and written information. Another possible use is the

transmission of cartoons. Pre-recorded audio cassettes could contain the story on one track and the pictures on the other.
Philips Telecommunicatie Nederland's experimental equipment included oscilloscopes for presenting the pictures. However, television screens are used in the current version. In addition, integrated circuits will be applied in future so that the equipment will then be smaller and cheaper. The main cost factor is the filter in the central unit which goes next to the telephone.

Philips Telecommunicatie, Netherlands.

(393 S)

On the trail of interference

The Automobile Equipment Group 1 of Robert Bosch GmbH has recently acquired a new weapon with which to hunt down annoying sources of radio interference. A computer-controlled VHF - UHF test assembly supplied by Rohde & Schwarz will be used for investigating interference suppression in cars, car parts of the company's own production — such as ignition systems, electric motors and generators —, film cameras marketed under the Bauer label and emergency generator units.
The main component of the test assembly, the VHF - UHF Test Receiver ESU 2, measures over the frequency range of 25 to 1000 MHz in compliance with the VDE guidelines 0875 to 0877 and 0879, the CISPR Publications 2, 4 and 12 to 14, and European Community directives. The Basic-programmable desktop computer 4051 from Tektronix presents the test results in graphic and tabular form on its large storage screen. Programmed evaluation of the results by the computer is also provided, as is permanent documentation on the Hard Copy Unit 4631.
This automated test assembly can be used for measurements on useful radio signals as well as on interference, and is suitable for operation in the field, in the laboratory or in a test vehicle. The assembly comprises the VHF - UHF Test Receiver ESU 2, the Frequency Controller EZK and two Code Converters PCW from Rohde & Schwarz, the Graphic Computing System 4051 and Hard Copy Unit 4631 from Tektronix, a digital voltmeter and a frequency counter.
For frequencies from 25 to 1000 MHz the receiver has a voltage measurement range of -10 to $+120$ dB (μV) and can measure two-port transfer constants from -90 to $+40$ dB.

(379 S)

SELEKTOR

Using heat to cool the Sahara

A small communications station in the north African desert is constantly heated up by the dissipation of its radio relay amplifiers. And yet even in the extreme midday heat the temperature inside the insulated shelter is about ten degrees below that outside. The cooling effect is produced without pumps and fans, entirely without the use of energy, by means of an ingenious 'rectifier' system of air and water cycles devised in the Siemens development laboratories. Working roughly on the principle of gravity-circulation hot-water heating, the system extracts heat from the interior during the day, stores it in a water tank and discharges it to the environment during the night. In this way the system acts as a 'refrigerator' during the day.

The cooling system was developed in an effort to protect communications equipment in extreme climates from excessively high temperatures, without additional energy having to be supplied for this purpose. The shelter (1.3 m³ volume), equipped with heating elements (150 W), fulfilled all expectations the very first time it was tested in the climatic chamber. The equally good results obtained in the test under extreme outdoor conditions encouraged Siemens to extend the tests to larger shelters with a volume of about 20 m³. The unpowered cooling system is based on the differences in density and weight of liquids caused by different temperatures. It consists of three cycles. The warm air rising from the equipment in the shelter is first guided by a baffle to heat exchanger A where it cools and consequently flows downwards again to the equipment. The second cycle is formed by the water in exchanger A. It heats up, expands and, as a result of its lower specific weight, passes through an ascending pipe to a central tank inside the shelter, while a downpipe with cooler water feeds exchanger A.

Heat exchanger B, which links the tank with the exterior, is warmer during the day, so that no cycle is created. At night, however, as soon as the outside temperature drops below the temperature of the tank, the third and most important cycle is started up: the water from exchanger B cools the contents of the tank — via an ascending pipe and a downpipe again — until the densities have become equal or until the ambient temperature rises above that of the tank again. The reservoir cooled in this way is then used for fully automatic cooling of the interior during the day.

Siemens AG, München, W-Germany.

(395 S)

Principle of the cooling system

I Inner water cycle	1 Downpipe to A
II Outer water cycle	2 Ascending pipe from A
A Heat exchanger, inside	3 Downpipe from B
B Heat exchanger, outside	4 Ascending pipe to B

'Charged' cars are no danger

You occasionally see them: cars dragging a 'tail' of metal braiding and hard rubber behind them, intended to discharge the static charge of the vehicle. The idea comes from the big tanker trucks, which always drag along a metal chain touching the ground. But does this 'discharge band' really serve any useful purpose? Engineers at the Siemens high-voltage laboratory in Berlin investigated to what extent cars are charged while being driven. Friction between the wheels and the road surface is the main cause for the electrical charge. The rubber-tired vehicle is electrically insulated as it rolls on the roadway; it is — from a physical point of view — a capacitor with a capacitance of about 100 pF. When the weather is dry and the insulation resistance of the tires is thus sufficiently high the 'autocapacitor' will be charged to approximately 10,000 V. The energy stored equals about 0.005 Ws. This amount of energy is so small that it can be discharged via the human body without causing any harm whatsoever. If a 'charged' car is touched with the hand, the energy will almost completely discharge in about a microsesond. In a darkened room this discharge could be perceived as very thin blue sparks, but in daylight or with street lighting it remains invisible. The whole effect is thus similar to the static charge experienced when one walks on synthetic carpets: it is a nuisance, but completely harmless.

The electrical voltage between the car body and the road surface can always be measured in terms of tens of thousands of volts even at high speeds since pointed or sharp-edged parts of the car body provide for local discharging and prevent higher voltages from being attained.

It is questionable whether conductive bands can prevent cars from becoming charged. The vehicle can only be charged when both the road and tires are dry — in the rain a good electrical discharge is always available. A dry road surface acts as a high electrical resistance, making it difficult to bring the charge from the car to the ground. For the owner of a passenger car there is really only one piece of advice: since it is impossible to protect oneself against spark discharges resulting from static charging it is better to accept them without a fuss. If one is prepared for the tingling sensation it is easier to take. People are not endangered by the charge on a car — regardless of whether or not one attaches a 'lightning conductor' to the car or not.

Siemens engineers once again advise motorists caught in a thunderstorm to drive to the nearest parking space and remain in the car. Since the vehicle is a closed Faraday cage passengers are best protected against lightning inside the car.

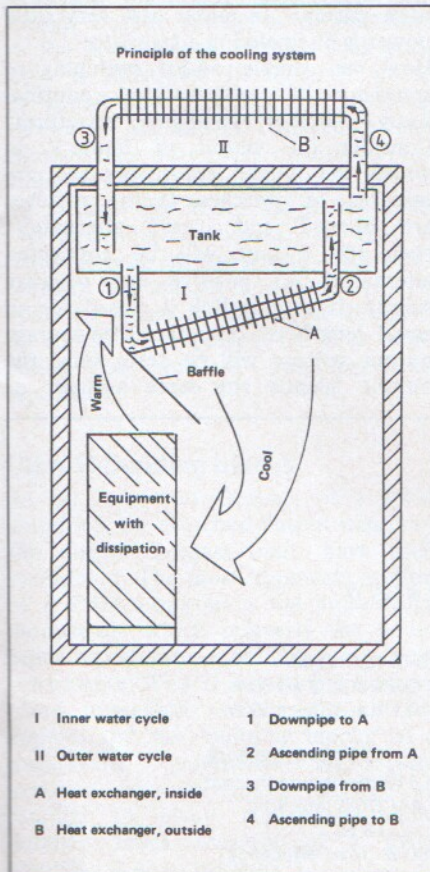*Siemens AG, München, W-Germany.*

(394 S)

PWM audio amplifiers

Although the concept of the pulse-width modulated or 'Class-D' audio amplifier has been known since the late nineteen-forties, there has been little practical exploitation of the technique apart from a few short-lived designs by Sinclair and Mullard in the nineteen-sixties.

However, recent advances in semiconductor technology, power FETs in particular, have made the class-D amplifier a more feasible proposition, and the PWM concept has been resurrected by some Japanese hi-fi firms.

This article takes a look at the principles of class-D amplifiers and examines the circuit of a commercial design.

The problem with conventional audio amplifiers, particularly when called upon to deliver large amounts of power, is that they are inefficient. This is a direct consequence of the fact that they operate in a linear fashion, i.e. the output voltage at any instant is equal to the input voltage multiplied by a constant — the voltage gain of the amplifier. Since the output voltage of an amplifier must be derived from the supply voltage via the output transistor(s) it follows that the difference between the output voltage and the supply voltage must be dropped across one of the output transistors, resulting in wasteful dissipation of energy as heat in the output transistors.

The output transistors in a linear amplifier can be likened to variable resistors whose resistance is adjusted by the input voltage so that the voltage developed across the amplifier load is always equal to the input voltage multiplied by the voltage gain of the amplifier. To take a simple example, suppose an amplifier is delivering a voltage equal to half supply into a load, R. The effective resistance of the output transistor is thus also R, and the same power is dissipated in both the load and the output transistor, representing an efficiency of only 50%.

The highest efficiency that a conventional class-B amplifier can hope to achieve is obtained when the peak output voltage (neglecting losses due to transistor saturation voltages) is equal to supply voltage, i.e. no voltage is dropped across the output transistors on signal peaks. For a sinewave input signal the efficiency achieved at this level is around 70%. Of course, in a practical situation with a music input the amplifier rarely achieves full output and the average efficiency is much lower.

Clearly the efficiency of a conventional amplifier is limited because the output transistors are used as 'rheostats' to regulate the voltage developed across the load. The only way that this can be avoided is to operate the output transistors as switches. This means that a transistor would either be turned hard on, passing current but with very little voltage developed across it, or cut off, in which case it would have full supply voltage across it but would be passing little current. In either case very little power is dissipated in a transistor.

How can simple on-off switching of transistors be translated into a continuously variable analogue waveform? Consider the circuit of figure 1, in which the output transistors are represented as switches. If the switches are opened and closed alternately, then the output will be connected alternately to positive and negative supply. If each switch is closed for an equal length of time then the average output voltage will be zero, since the output spends the same amount of
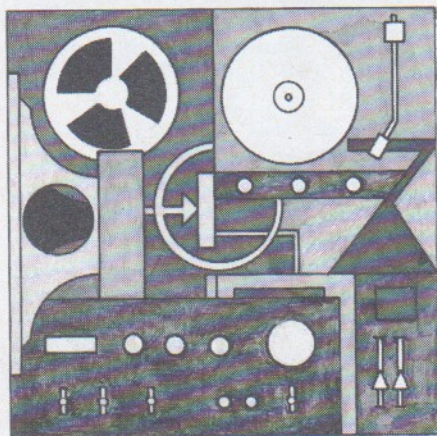
Table 1.

Specifications of the TA-N88

Output power: (both channels driven simultaneously)	2 x 160 W
Damping factor:	20 (8 Ω, 1 kHz)
Harmonic distortion:	≤ 0.5% at ≤ 2 x 160 W
IM distortion: (60 Hz/7 kHz; 4:1)	≤ 0.1% at ≤ 2 x 160 W
Frequency response:	5 . . . 40,000 Hz ± 0.5 dB
Signal-to-noise ratio:	≥ 110 dB (inputs shorted)
Input sensitivity:	1.4 V RMS into 50 k
Loudspeaker impedance:	8 . . . 16 Ω
Total consumption under full load:	550 W (total efficiency over 58%)
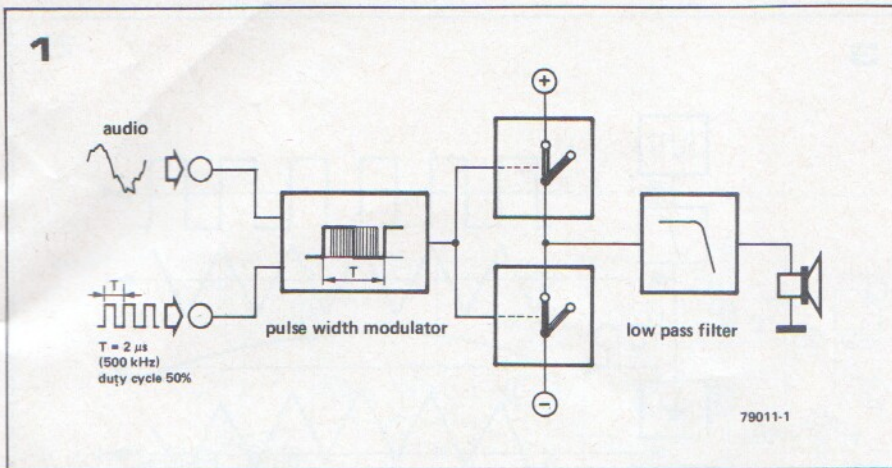
time at the positive potential as it does at the negative potential.

If the upper switch is closed for a longer time than the lower switch, i.e. the duty-cycle is greater than 50% then the average output voltage will no longer be zero but will be positive. Conversely, if the duty-cycle is less than 50% then the average output voltage will be negative. Since one output transistor is always turned hard on and the other is turned off, there is little power dissipated in the output transistors.

These are the essential principles of a PWM amplifier. The output transistors switch the load alternately to the positive and negative supply at a rate much greater than the highest audio frequency. The audio input signal is used to control the duty-cycle of the switching waveform such that the average output voltage is proportional to the audio input voltage. This principle is illustrated in figure 2, where 23 consecutive cycles of the switching waveform are shown underneath one another. The duty-cycle modulation for a sinewave input is clearly visible.

To retrieve the audio signal free from the switching waveform, all that is required is to interpose a lowpass filter between the output and the load. To minimise losses in the filter it must, of course be a passive LC filter consisting of low-loss inductors and capacitors.

Class-D building blocks

So far only two sections of a PWM amplifier, the switched output stage and the output lowpass filter, have been considered. The most important section of a PWM amplifier is the pulse-width modulator, which converts the audio input signal into a variable duty-cycle squarewave. This is not such a complicated procedure as it might at first appear, and the 'building blocks' of a pulse-width modulator are quite standard circuits which should be familiar to many readers. A more detailed block diagram of a PWM amplifier is given in figure 3.

The first step is to generate a squarewave with a 50% duty-cycle at the
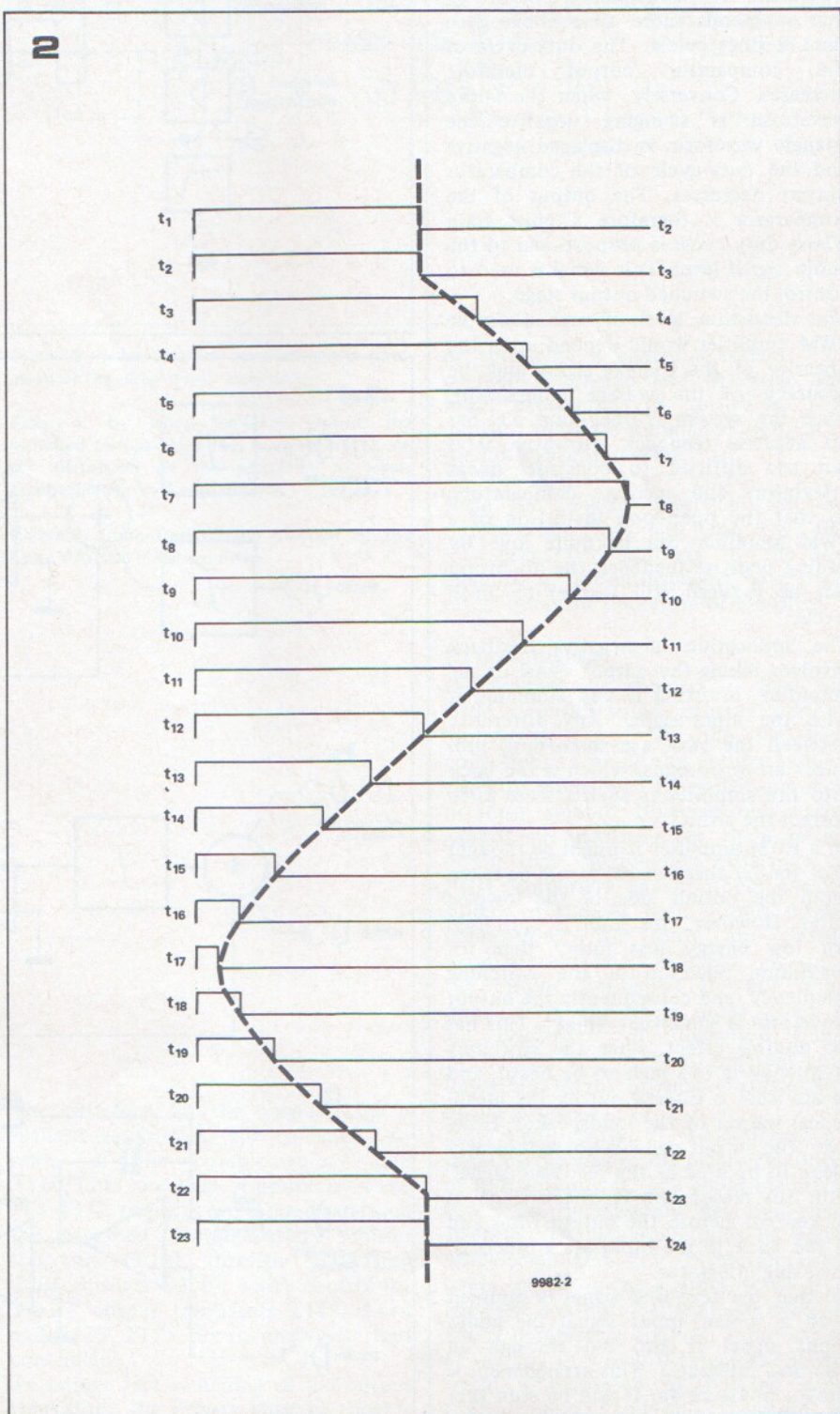
switching frequency. This is then fed through an integrator to produce a triangular waveform of the same frequency.

The triangle waveform is summed with the audio waveform and the resultant signal is fed to a zero-crossing detector, i.e. an analogue voltage comparator which has one input tied to zero volts. If the audio signal level is zero then the signal fed to the comparator is simply the triangle waveform whose zero-crossing points occur precisely in the centre of the waveform, so the output of the comparator is a squarewave with a 50% duty-cycle. If the audio waveform is going positive then the triangle waveform is also displaced positive so that it spends more time above zero than it does below. The duty-cycle of the comparator output therefore increases. Conversely, when the audio waveform is swinging negative the triangle waveform is displaced negative and the duty-cycle of the comparator output decreases. The output of the comparator is therefore a pulse train whose duty-cycle is proportional to the audio signal level. This signal is used to control the switched output stage.

The distortion level of such a simple PWM amplifier would depend upon the linearity of the triangle signal and the accuracy of the voltage comparator, since the system is open-loop, i.e. has no negative feedback. However, it is not too difficult to construct linear integrators and accurate comparators, so that the open-loop distortion of a PWM amplifier can be quite low. By adding negative feedback the distortion can be reduced still further to 'hi-fi' levels.

The application of negative feedback involves taking the output signal of the amplifier, inverting it and summing it with the input signal. Any difference between the two, e.g. distortion, produces an error signal which is fed back into the amplifier in such a sense as to correct the error.

In a PWM amplifier it might be thought that the feedback signal would be taken from the output side of the lowpass filter. However, this filter is optimised for low energy loss rather than for maximum rejection of the switching frequency, and consequently the output waveform is somewhat 'spiky'. This has no audible effect, since the switching frequency is too high to be heard, and in any case is filtered out by the mechanical inertia of the loudspeaker. However, the output signal is not sufficiently clean to be used as the feedback signal. For this reason the feedback signal is picked off before the output filter and is fed back to the input via a precision inverting integrator.

So that the feedback signal is summed with a similar input signal the audio input signal is also fed through an identical integrator. This arrangement is shown in figure 4a. It will be seen that three integrators are involved, one for
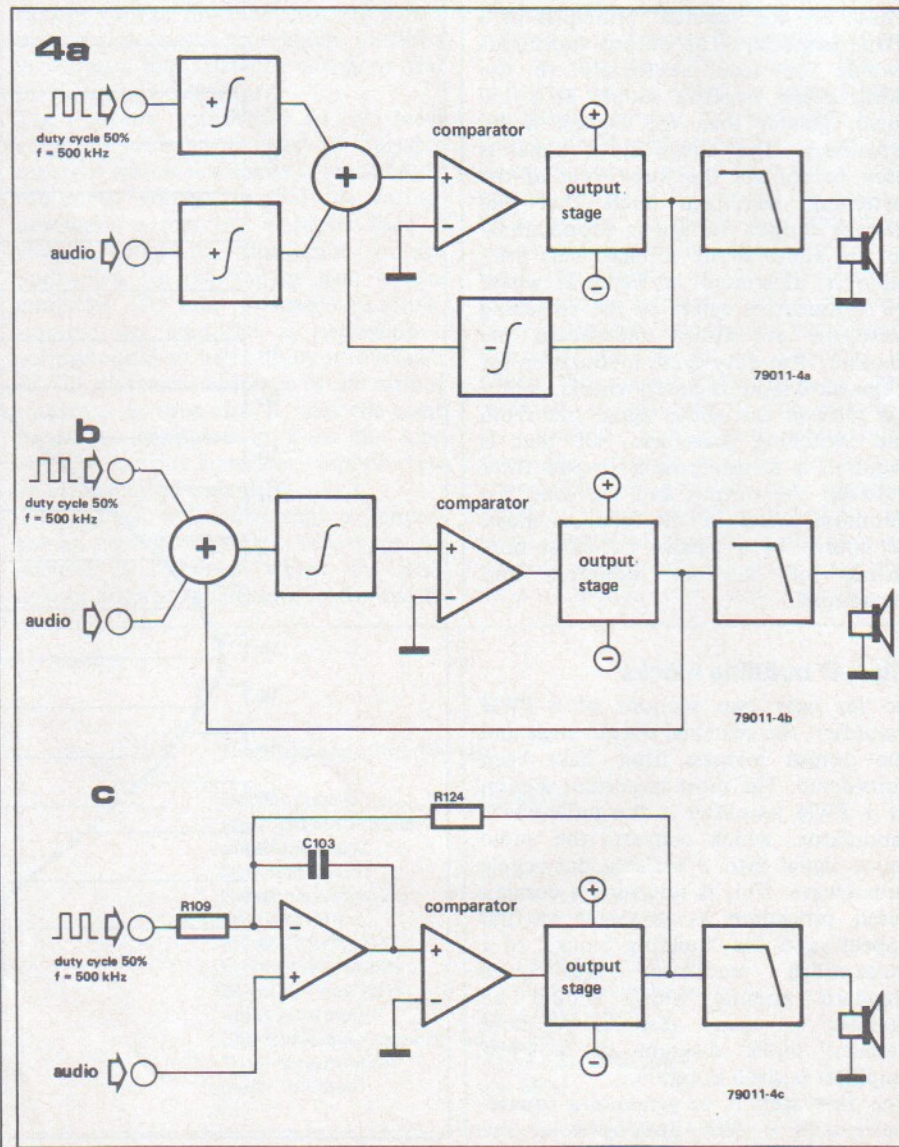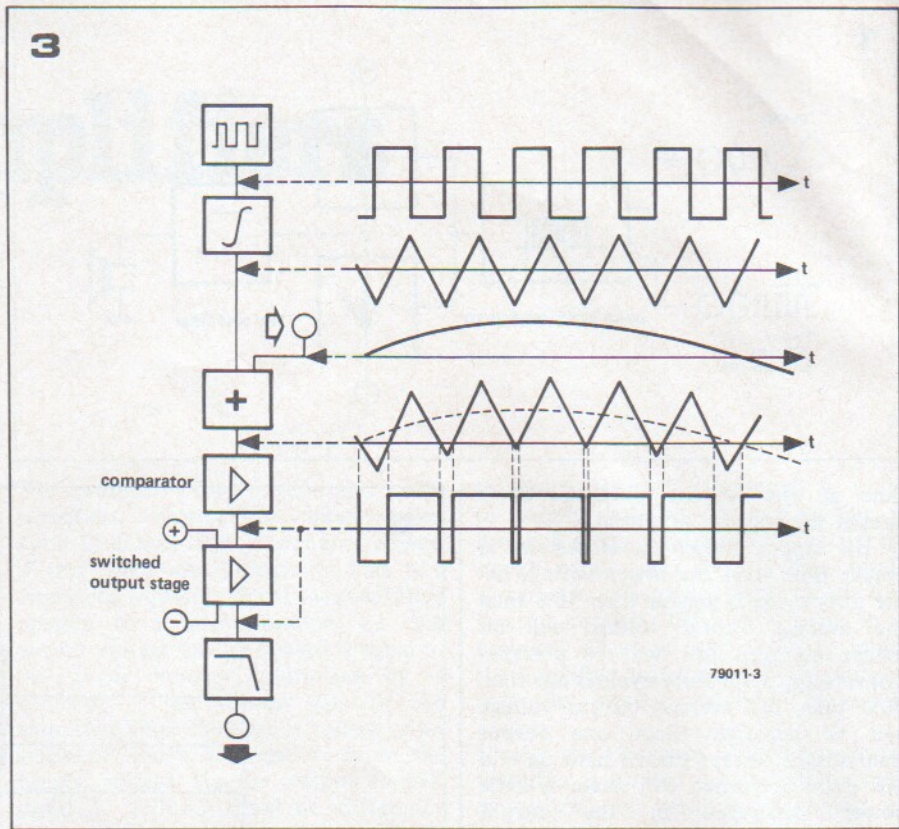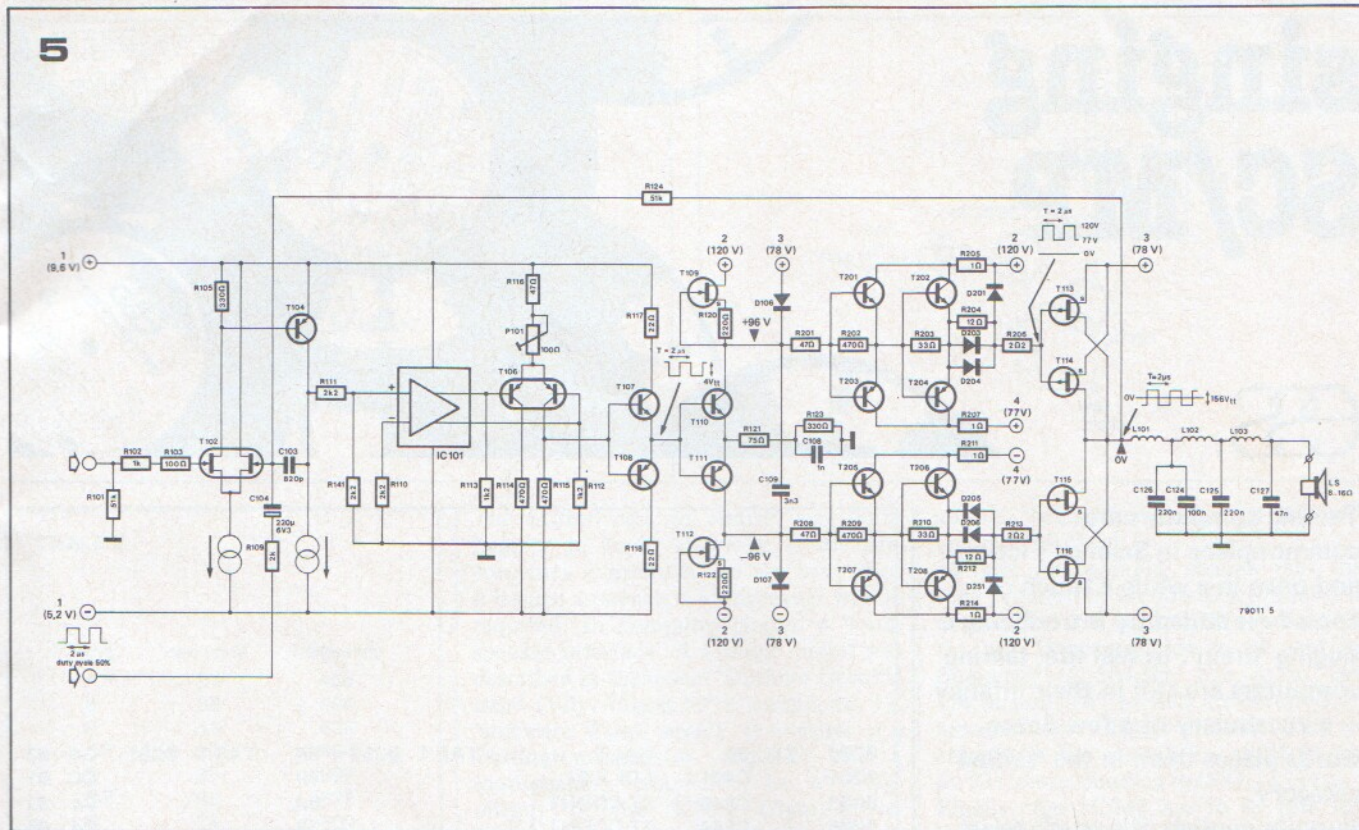
79011-3

79011-4a

79011-4b

79011-4c

5

the feedback signal, one for the input signal, and a third to convert the square-wave signal into a triangle waveform. A more economical system is to use a single integrator placed after the summing point, as shown in figure 4b. If this circuit is elaborated slightly, as shown in figure 4c, we in fact obtain the block diagram of a currently available commercial PWM amplifier — the TA-N88 from Sony.

From theory to practice

The basic circuit diagram of the TA-N88 is shown in figure 5. The circuit shown does not include the power supply, which is also pulsed (at 20 kHz), or the current limiting (if excessively high currents appear on the +3 or −3 V supply lines, the input voltage at the junction of R102/R103 is shorted to ground), thermal overload and DC protection circuits (in the latter case this is simply a relay in the loudspeaker lead). Furthermore, the 500 kHz oscillator and buffer stage are also omitted. The integrator of figure 4c is formed by differential amplifier T102 (this is a dual transistor which determines the DC offset at the output of the amplifier) plus T104; both transistors are connected as a constant current source. The squarewave output from the oscillator is introduced via R109; C103 is the integrator capacitor.

The comparator is formed by four differential amplifiers connected in cascade. Three of these are contained in IC101, whilst the fourth is provided by the dual transistor T106. The comparator output is buffered by the complementary emitter follower T107/T108.

Figure 3. Pulse diagram illustrating the operation of the pulse width modulator.

Figure 4. By using feedback around the switched output stage it is possible to reduce the distortion of the amplifier to levels acceptable for hi-fi applications.

Figure 5. Simplified circuit diagram of the Sony TA-N88 PWM amplifier.

The switching for the output stage is realised as follows: the collector currents of the complementary pair T110/T111 are held constant by T109 and T112 respectively. This means that the cascaded complementary emitter followers (T201 through T204 and T205 through T208), which control the V-FET output transistors T113, T114 and T115, T116, are themselves current controlled.

To ensure fast switching of the output transistors, the drivers must be capable of supplying a generous amount of current to charge and discharge the input capacitance of the FETs. This explains the large number of transistors used in the output stage.

L101 . . . L103 and C124 . . . C127 form the lowpass filter between the output transistors and load (i.e. the loudspeaker). Variations in the impedance of the loudspeaker have little effect upon the operation of the filter.

In conclusion

Cynics may be inclined to view the PWM amplifier as yet another spurious 'innovation' sold by the big hi-fi companies to an all too gullible public, eager to latch onto the latest gimmick or status symbol. However such a view cannot be justified, for there is no doubt that the inherent efficiency of PWM amplifiers makes them ideally suited to today's high output power applications. Not only are class-D amplifiers extremely efficient, but, since the output transistors are not being operated in the linear mode, there is no possibility of distortion being introduced as a result of inherent non-linearities in the transfer characteristic of the devices.

More significantly, the TA-N88 may be pointing the way to the future evolution of hi-fi equipment. The audio manufacturers are increasingly devoting their attention to digital methods of signal processing, and the PWM amplifier could prove to be a transitional step between conventional analogue systems and a completely digitised audio chain, at the end of which the analogue music signal is recovered by nothing more than a couple of coils and capacitors!

singing SC/MP

ess

Talking computers are commonplace in Science Fiction, and once in a while a touch of comedy is added by introducing a singing 'brain'. In real life, talking computers are still in their infancy — a vocabulary of a few dozen words places them in the 'skilled' category.

Funnily enough, it's much easier to make an 'electronic brain' hum a tune! Even the relatively slow and dull-witted microprocessor can reproduce melodies with a high degree of accuracy. This is demonstrated in the SC/MP program described here. With Christmas in the air, a well-known tune was chosen: 'Silent Night'.

Programs for several other well-known tunes are also included on an ESS record. For interested readers, who are not (yet) the proud owners of a SC/MP system, the μP's rendition of these tunes is recorded 'live' on the B-side of the same disc.

TABLE 1

Addr	Label	Hex	Mnemonic	TAB 1 (0C53)	TAB 2 (0C81)
0C00	$ 1	00	HALT	5A	C4 03
0C01		C401	LDI X'01	49	DC 01
0C03		C84B	ST COUNT 1	5A	C4 02
0C05		C84A	ST COUNT 2	78	C4 05
0C07		C453	LDI X'53	5A	C4 03
0C09		31	XPAL 1	49	DC 01
0C0A		C40C	LDI X'0C	5A	C4 02
0C0C		35	XPAH 1	78	C4 05
0C0D		C481	LDI X'81	25	EB 05
0C0F		32	XPAL 2	25	C4 03
0C10		C40C	LDI X'0C	39	D6 07
0C12		36	XPAH 2	32	83 08
0C13		C42F	LDI X'2F	32	57 06
0C15		C83B	ST COUNT 3	5A	EB 05
0C17	$ 3	C418	LDI X'18	49	DC 04
0C19		8F00	DLY X'00	49	DC 02
0C1B	$ 2	C402	LDI X'02	32	9D 05
0C1D		07	CAS	39	F7 01
0C1E		C400	LDI X'00	49	DC 02
0C20		8F00	DLY X'00	5A	C4 03
0C22		C400	LDI X'00	49	DC 01
0C24		07	CAS	5A	C4 02
0C25		C400	LDI X'00	78	C4 05
0C27		8F00	DLY X'00	49	DC 04
0C29		B825	DLD COUNT 1	49	DC 02
0C2B		9CEA	JNZ $ 3	32	9D 05
0C3D		C024	LD COUNT 1'	39	F7 01
0C2F		C81F	ST COUNT 1	49	DC 02
0C31		B81E	DLD COUNT 2	5A	C4 03
0C33		9CE6	JNZ $ 2	49	DC 01
0C35		08	NOP	5A	C4 02
0C36		B81A	DLD COUNT 3	78	C4 05
0C38		98C6	JZ $ 1	25	EB 05
0C3A		C501	LD@ 1 (1)	25	C4 03
0C3C		C8E9	ST DELAY	14	83 08
0C3E		02	CCL	25	93 02
0C3F		F432	ADI X'32	39	F7 02
0C41		C8DD	ST DELAY	32	9D 0A
0C43		C601	LD@ 1 (2)	19	F7 08
0C45		C809	ST COUNT 1	32	9D 05
0C47		C80A	ST COUNT 1'	5A	C4 01
0C49		C601	LD@ 1 (2)	78	6E 03
0C4B		C804	ST COUNT 2	5A	C4 05
0C4D		90CC	JMP $ 1	6D	AF 01
0C4F		00	4 counter bytes	8F	93 02
0C50		00		A9	6D 0C
0C51		00			
0C52		00			

1

2

TABLE 2

Note	hex code	frequency
A#	C7	466
B	B8	494
C	A9	523
C#	9C	554
D	8F	587
D#	83	622
E	78	659
F	6D	698
F#	63	740
G	5A	784
G#	51	831
A	49	880
A#	41	932
B	39	988
C	32	1047
C#	2B	1109
D	25	1175
D#	1F	1245
E	19	1319
F	14	1397
F#	ØF	1480
G	ØA	1568
G#	Ø6	1661
A	Ø2	1760

Figure 1. Only three components are required for a 'loudspeaker interface'.

Figure 2. In case of availability problems, the BC 517 can, of course, be replaced by a standard Darlington configuration.

Table 1. Complete program listing for 'Silent Night'.

Table 2. Using the information given in this Table, it is a fairly easy matter to program the SC/MP for other melodies.

The SC/MP can, of course, sing quite happily to itself. However, if the performance is intended for the benefit of a human audience, a loudspeaker will be required. In computer jargon: a 'loudspeaker interface'. A suitable circuit was described in the recent 'Summer circuits' issue (July/August 1978, circuit no. 12, 'software Kojak siren'). It consists of a Darlington-transistor amplifier and a loudspeaker, connected to a 'Flag' output of the SC/MP (figure 1; an alternative circuit is given in figure 2). If the Flag is set and reset rapidly, a tone is produced; the more rapidly the Flag changes state, the higher the output frequency.

A melody consists of a succession of 'tones' with different frequencies. To make the SC/MP 'sing', it must be programmed to set and reset one of its Flags at a frequency that is determined by a list of numbers somewhere in its memory. Furthermore it must be told, by means of a second list of numbers, how long each note should last. In other words, a program is required that will combine two lists of numbers (one for tone pitch and one for tone duration) to produce a melody. A suitable program is given in Table 1.

The output frequency generated at any given moment is determined by a hexadecimal number XX as follows:

$$f = \frac{10^6}{556 + 8(XX)_{16}}\ Hz,$$

where XX is limited to the range $0 \leqslant XX \leqslant CD$ (hexadecimal!).

This limits the SC/MP's singing to two octaves, as shown in Table 2. To avoid the need for combersome calculations, this Table lists the possible notes, the corresponding hexadecimal numbers and the output frequencies. With this information, it is a relatively simple matter to draw up the first list (giving the frequencies of the consecutive notes) for any given melody.

As stated earlier, the duration of each note is determined by a number in a second list. In this program, the tone duration is entered as the number of periods the note is to last. Since this can lead to fairly large numbers, two bytes are reserved for each note, the total number being the *product* of the two numbers involved.

To give an example, assume that the tone required is the low D and that it is to last for 1/4 second. From Table 2, the low D corresponds to 8F and its frequency is 587 Hz. For it to last 1/4 second, 147 periods are required. In hexadecimal: $(93)_{16}$ periods. This is entered in the second list as 93 Ø1 (or Ø1 93), corresponding to $(93)_{16}$ x $(1)_{16}$.

Finally, the SC/MP has to be told the length of the tune, i.e. the number of notes. Or, to be more precise, the number of notes plus one. This value is entered in address ØC14.

In the program, TAB 1 (starting at position ØC53) is the list of notes required, and TAB 2 is the list of tone durations. The start address of the latter list is stored in ØCØE (lower address byte) and ØC11 (higher address byte).

As an example, the complete program listing for 'Silent Night' is given in Table 1. It is entered from address ØCØØ; this is also the start address. The program is started by operating the halt/reset key.

If other melodies are to be produced, the lists of numbers under TAB 1 and TAB 2 must be modified accordingly. Furthermore, as mentioned earlier, the start address of TAB 2 must be entered in positions ØCØE and ØC11, and the number of notes plus one is stored in ØC14.

The program described here is included, with 5 other well-known Christmas melodies and 'Mary had a little lamb', on the Elektor Software Service record ESS 002. The B-side of the same recording contains the SC/MP's 'real-time' rendition of the same tunes.　◄

Litt: Experimenting with the SC/MP, Nov. 1977 . . . April 1978;
Software Kojak siren, July/August 1978.

tone-burst generator

A (repetitive) tone-burst signal is an extremely useful aid for testing audio equipment. Basically, this type of test signal is obtained by switching the output of a sine-wave oscillator on and off at regular intervals. The generator described in this article utilises a novel design approach that simplifies the circuit considerably and only involves one minor reduction of the capabilities.

The sinewave is the most commonly used test signal. It is simple to analyse (both 'on sight' and mathematically) so that any distortion can usually be recognised quickly. Its very simplicity, however, is also its main disadvantage: it has very little in common with the signals that an audio system is normally expected to handle: music and speech.

Audio signals are extremely 'dynamic': transients and other more-or-less rapid changes in level are actually the most important information in a speech signal. In order to test a system which is intended to handle this type of material, it seems reasonable to look for a 'dynamic' test signal. There is no way to measure 'transient response' with a signal that is as obstinately steady-state as DC. And that, regrettably, applies to sinewaves.
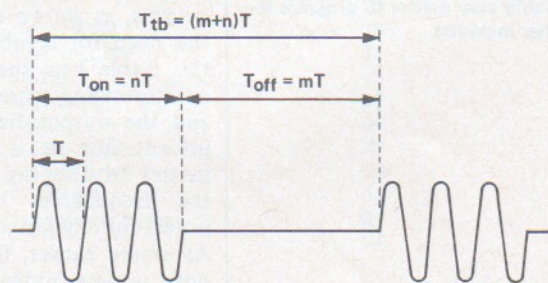
What about squarewaves? A good second in the list of commonly used test signals. It is definitely better than the sinewave when it comes to pointing out poor transient response. However, it is definitely inferior on several other counts. Just think: a digital NAND gate will pass a squarewave beautifully — but a NAND gate makes a very poor audio amplifier indeed . . .

A tone-burst signal can be considered as a combination of sinewave plus squarewave. It has the advantages of both: it is steady-state for a while, then suddenly changes to a new 'steady state' and so on. A typical tone-burst signal is shown in figure 1. It consists of one or more sinewaves, then a gap (one or more sinewaves in length), then again one or more sinewaves and so on. That it is in some ways similar to a sinewave is obvious; the similarity with a squarewave is perhaps less apparent, until one realises that it is basically equivalent to the output from a sinewave generator that is being turned on and off by a squarewave generator of lower frequency.

All very well, but how does one obtain a 'tone-burst' signal? Apparently, a squarewave generator must 'gate' the output from a sinewave generator. One way of achieving this is shown in figure 2. The sinewave is fed to an electronic switch. As the switch is opened and closed, a succession of sinewave 'bursts' will appear at the output. Control of the switch is rather complicated (more than this simplified block diagram suggests!). The sinewave is fed to a zero-crossing detector; the output from the latter is used as 'clock' signal for two programmable counters. Only one of these counters is 'active' at any given moment — the other is held in the 'reset' position by a flip-flop. When the selected maximum count of the 'active' counter is reached, the flip-flop is triggered. The first counter is then reset and the other counter is enabled. Since the output of the flip-flop also drives the electronic switch, the final result is a number of sinewave periods determined by one counter, followed by a 'dead
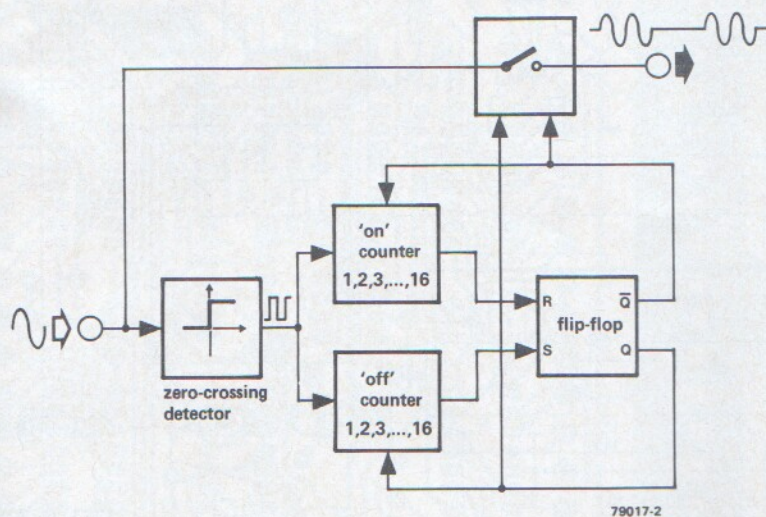

1

$T_{tb} = (m+n)T$

$T_{on} = nT$

$T_{off} = mT$

T

$n = 1,2,3,...,16$
$m = 1,2,3,...,16$

$\dfrac{T_{tb}}{T} = 2,3,4,...,32$

79017-1

2

79017-2

3

79017-3

that matter, they are good enough for professional use. The only problem is that they make for less sweeping advertisements . . .
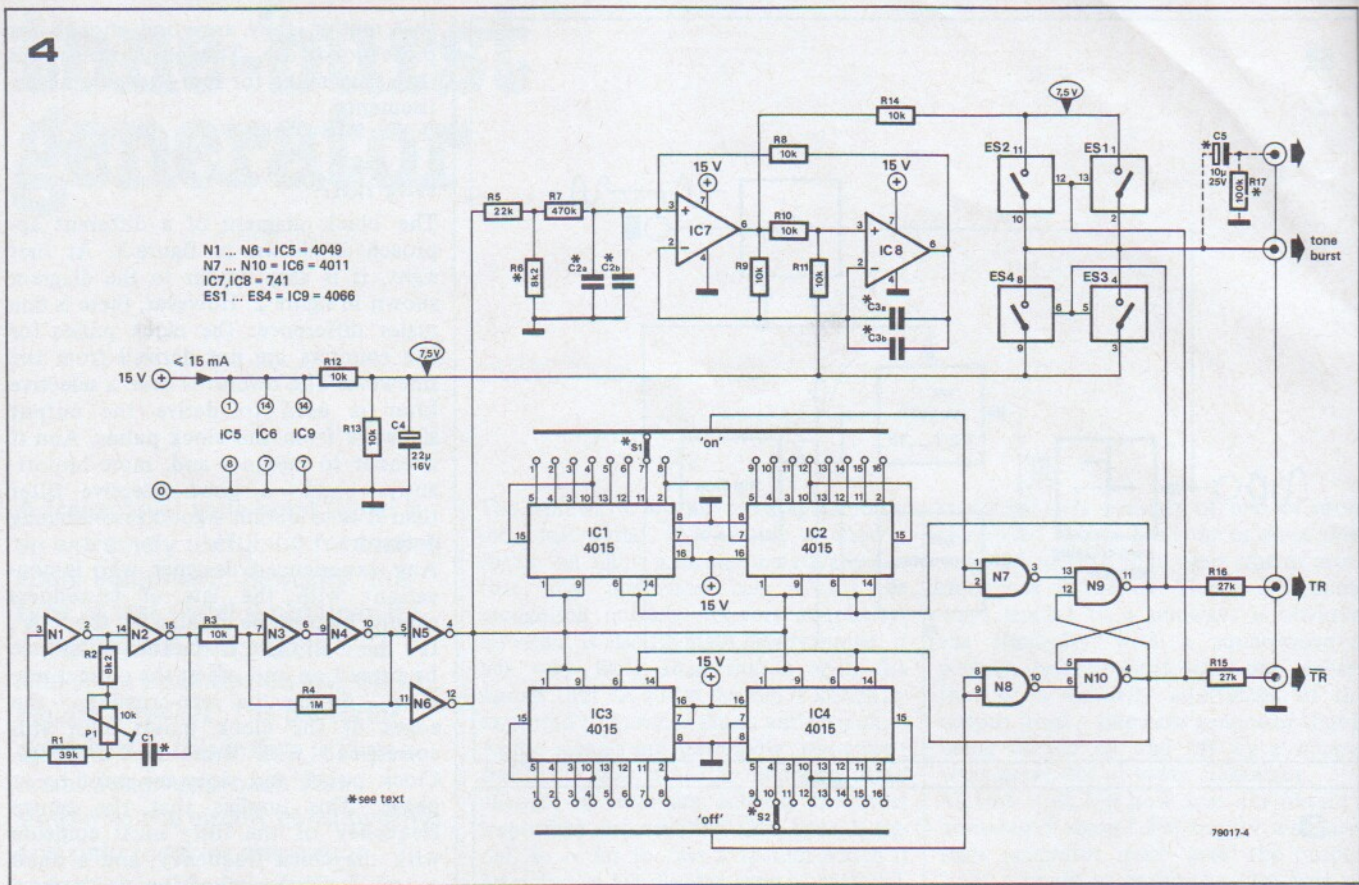
Why not?

The block diagram of a different approach is shown in figure 3. At first sight, it is very similar to the diagram shown in figure 2. However, there is one major difference: the clock pulses for the counters are not derived from the sinewave. The reverse is true: a selective filter is used to derive the output sinewave from the clock pulses. And it is easier to design — and, more importantly, build — a good selective filter than it is to obtain a good zero-crossing detector.

Any experienced designer who is conversant with the law of cussedness should, by now, be looking for the 'bug'. It's there all right. If the tone-burst is to be turned on and off at the correct moment — during the zero-crossings — the edges of the clock pulses must still correspond with those zero-crossings. Clock pulses and sinewave must be in phase. This implies that the centre frequency of the filter must coincide with the clock frequency, and a fixed centre frequency therefore leads to a fixed clock frequency. Tone-bursts with 'swept' sinewave frequency are no longer possible. So what . . . who needs 'em, anyway?

The circuit

The complete circuit of the tone-burst generator is shown in figure 4. The burst length can be set, by means of S1, at anything between 1 and 16 complete sinewave periods. The interval between bursts is selected in the same way by S2. The clock generator, N1/N2, is a fairly standard circuit. Its output is not particularly 'clean', but suitable processing by the other four inverters in the same package (N3 . . . N6) produces a good squarewave. As illustrated in the block diagram (figure 3), this signal is fed to two counters, one of which (IC1/IC2) determines the length of the burst while the other (IC3/IC4) fixes the interval between bursts. One output of each counter is selected by S1 and S2, respectively, and used to set and reset the flip-flop (N9/N10). The Q and $\bar{Q}$ outputs of the flip-flop are fed back to the counters in such a way that when the count selected by S1 is reached, toggling the flip-flop, the corresponding counter is reset and the other is enabled. In this way, the two counters are used alternately.

The flip-flop outputs are also used to operate the electronic switches S1 . . . S4. When S1 and S2 are closed, the sinewave appears at the output; opening S1 and S2 and closing S3 and S4 blocks the sinewave and passes a DC level, corresponding to the zero level of the sinewave, instead.

zone' (no output) determined by the second counter.

A fool-proof system, one would think. However, there is at least one weak link in the chain: the zero-crossing detector. If the output tone-bursts are to start and stop in the zero-crossings of the sinewave, an accurate zero-crossing detector is required — not to mention zero phase-shift throughout the complete chain from detector through counters and flip-flop to switch.

These problems can be solved — witness the proliferation of commercial tone-burst generators that work according to this principle. However, why bother? A different approach obviates the whole problem. The results are certainly good enough for the home constructor — for

Figure 1. A tone-burst is basically equivalent to a sinewave that is switched on and off at regular intervals. Both the length of the burst and the period between bursts are a whole multiple of the sinewave period.

Figure 2. Block diagram of a conventional tone-burst generator.

Figure 3. Block diagram of a novel alternative.

4

The sinewave is derived from the clock pulses, as described earlier. The clock pulses are passed through a selective filter (IC7/IC8); the centre frequency of this filter, which corresponds to the clock frequency, is also equal to the sinewave frequency of the tone-burst. If a different tone-burst frequency is required, the clock frequency must be altered by choosing a different value for C1, and the centre frequency of the selective filter must be modified by altering the values of C2a/b and C3a/b.

Construction and calibration

A suitable printed circuit board design is shown in figure 5. Fairly standard components are used throughout, with the possible exception of the switches S1 and S2. Several alternatives are possible at this point, depending on availability and cost: 16 single-pole single-throw switches; a 24-way switch (more common than 16-way); a 12-way switch (four outputs of each counter are left unused); two or more switches in cascade; or even hard-wired internal programming of one useful combination (four periods on, say, and eight periods off).

The amplitude of the sine-wave bursts is approximately 8 V peak-to-peak. A 7.5 V DC component is also present in the output signal - corresponding, of course, to the 'zero' level between bursts. This DC component can be blocked by adding C5 and R17, as shown in dotted lines in the circuit diagram. The amplitude of the output sinewave can be modified by altering the value of R6.

Two 'trigger' outputs, TR and $\overline{TR}$, are also provided. They are derived from the outputs of the flip-flop, so that they change state at the beginning and end of each burst and can therefore be used to trigger an oscilloscope, giving a stable display.

The desired sinewave frequency determines the value of three capacitors: C1, C2 and C3. If the frequency, f, is given in kHz then the values of the capacitors in nF can be found as follows:

$$C2 = C3 = \frac{16}{f};$$

$$C1 = \frac{33}{f}.$$

The values of C2 and C3 are fairly critical, and two positions are reserved on the p.c. board for each of these capacitors so that the desired value can be approximated fairly accurately by connecting two capacitors in parallel. For instance, if an exact 1 kHz sinewave is desired, C2 and C3 would have to be 16 nF; this can be obtained by connecting a 15 n and a 1 n capacitor in parallel. The value of C1 is not so critical, since the frequency of the clock generator can be set correctly by means of P1.

The highest frequency obtainable is 20 kHz.

The current consumption of the circuit is quite low: 12 ... 15 mA.

The calibration procedure is extremely simple — only one potentiometer (P1) requires attention. The idea is that the clock frequency must coincide exactly with the centre frequency of the active filter, as otherwise the toneburst will not start and stop at the zero-crossing of the sinewave. The adjustment can be carried out quite easily with the aid of an oscilloscope. Photos 4 and 5 illustrate two incorrect settings; correct adjustment will produce the result shown in figure 6.

Figure 4. Complete circuit for a tone-burst generator that is more suitable for home construction.

Figure 5. Printed circuit board and component layout for the tone-burst generator (EPS 79017).

Photo 1. Tone burst, 1 sinewave period with a 16-period gap.

Photo 2. Tone burst, 16 sinewave periods with a 1-period gap.

Photo 3. Tone burst, 16 sinewave periods with a 16-period gap.

Photos 4 and 5. Clock signal and tone-burst with P1 incorrectly adjusted: the burst does not start and stop in the zero-crossings of the sinewave.

Photo 6. Tone-burst output after P1 has been accurately adjusted.

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79017
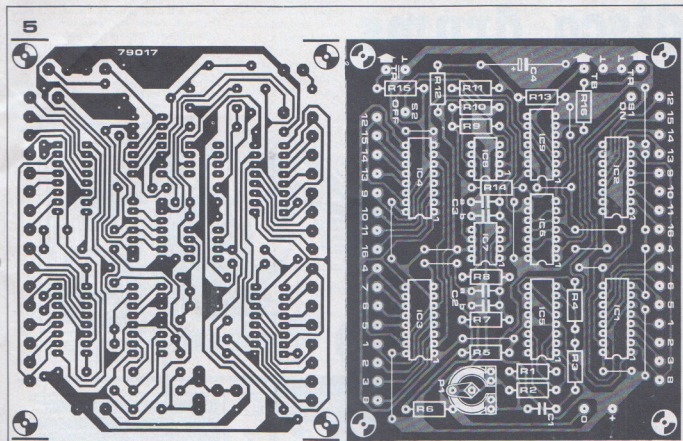
Parts list

Resistors:
R1 = 39 k
R2,R6* = 8k2
R3,R8 ... R14 = 10 k
R4 = 1 M
R5 = 22 k
R7 = 470 k
R15,R16 = 27 k
R17* = 100 k
P1 = 10 k preset

Capacitors:
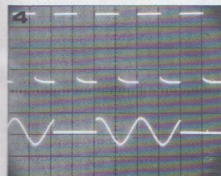C1* = 33 n
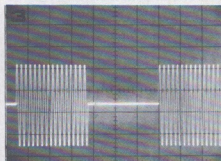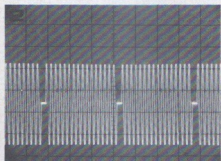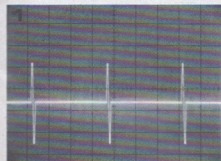C2a/b*, C3a/b* = 15 n + 1 n
C4 = 22 µ/16 V
C5* = 10 µ/25 V

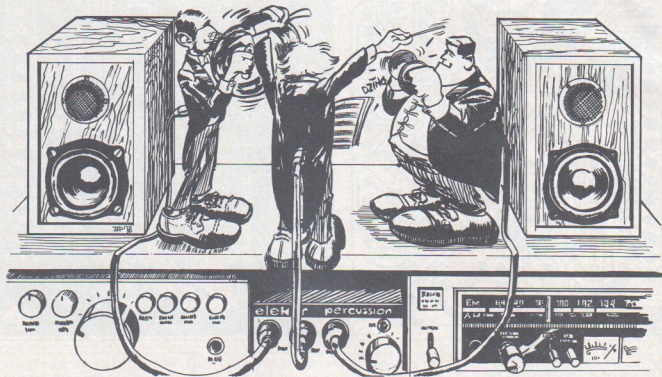**Miscellaneous:**
S1,S2 = single pole, 16-way
switch*

Semiconductors:
IC1 ... IC4 = CD 4015
IC5 = CD 4049
IC6 = CD 4011
IC7,IC8 = 741
IC9 = CD 4066

* see text

disco drums

A large number of contemporary pop records are aimed at the so-called 'disco scene'. A distinctive feature of most disco music is a heavy, repetitive and fairly complex percussive sound, which often has a characteristically artificial or 'funky' quality. With the aid of the circuit described here, records from the 'pre-disco era' or one's party tapes can be lent this authentic disco 'flavour' by the addition of a synthesised percussion track.

L. Witkam

It goes without saying that the 'drumbox' must have some means of detecting the beat of the piece of music being played, in order that the percussion be in time with the music. The simplest way of doing this is to utilise the fact that, almost invariably, the bass line in pop music lays down the beat for the rest of the instruments. Furthermore, the bass part tends to heavily accentuate the basic beat, making it easy to detect and follow. Thus the circuit operates by monitoring the level of the low frequency signal components and adding in the desired disco sounds at the appropriate points.

The drum sound is produced by a noise signal with a typically percussive attack-decay envelope. That is to say, initially the amplitude of the noise signal rises sharply to its peak value, then is made to die away more slowly (exponentially).

The block diagram of the drumbox is shown in figure 1. As can be seen, the music signal is first fed to a lowpass filter which eliminates all but the bass frequencies. The resulting signal is then rectified and used to control an opto-coupler. When the latter is actuated, a noise signal is fed to the summing circuit, where it is mixed with the original music signal.

Circuit diagram

The complete circuit diagram of the unit is shown in figure 2. The lowpass filter is formed by T1 and T2. The filter slope is 18 dB per octave (i.e. the filter is third-order) and the turnover frequency is 40 Hz. The filter output is rectified by diodes D1 and D2, and smoothed by capacitor C7. This capacitor, which has a very short charge time, stores the instantaneous value of the bass part of the music signal. Thus when the voltage across C7 is sufficiently large, T3 is turned on. C7 then discharges via R7 and the base-emitter junction of T3, until, after a short time, this transistor is once more turned off. However, during the time that T3 remains turned on both LED D3 and the opto-coupler LED are lit, thereby turning on the opto-coupler transistor. This means that the noise signal, which is applied to the base of this transistor, is fed from its emitter to T4, where it is amplified before being mixed with the original music signal via R17.

The noise signal itself is produced by transistor T6, which is connected as a zener diode, and is amplified by T5. The volume of the noise signal can be varied by means of P3, whilst P1 alters the sensitivity of the circuit (i.e. the level of the bass part at which a 'drumbeat' is produced). P2 should be adjusted until a suitable noise signal is obtained. Varying the position of this preset will have some effect upon the timbre of the noise; one therefore has a certain element of choice regarding the type of sound produced. The noise level will also vary considerably depending upon

Figure 1. Block diagram of the 'disco drumbox'. Whenever the circuit detects a pronounced bass note in the passage of music, it adds the desired percussive effect in the form of a noise signal.

Figure 2. Complete circuit diagram. The operation of the circuit can be checked by testing the voltages at the points indicated. The values given in the table apply to quiescent conditions.

the sample of transistor used. Thus it is recommended that the noise generator be built first and tested using different transistors.

Construction of the circuit should not present any special problems. A T111 or T112 from Texas Instruments can be used for the opto-coupler. Both types are available in a six-pin DIL package. The pinout details given in the circuit diagram apply to both versions.

The quiescent current consumption of the circuit is roughly 3 mA; maximum consumption (during each 'beat') is

approximately 40 mA, which means that, if desired, the unit can be powered by battery. The operation of the circuit can be checked with the aid of the measurement points and test voltages indicated in the circuit diagram.

Although in theory the unit can be inserted at any point in the (pre-) amplifier chain, it is recommended that it be placed before the volume control of the amplifier. The reason for this is that otherwise it would be necessary to alter the setting of P1 every time one adjusted the volume control. ◀◀

79001 1

T1 ... T6 = BC 549B
IC1 = TIL 111, TIL 112

|	
①	0,8 V
②	6,3 V
③	9,0 V
④	0 V
⑤	5,6 V
⑥	5,6 V

79001 - 2
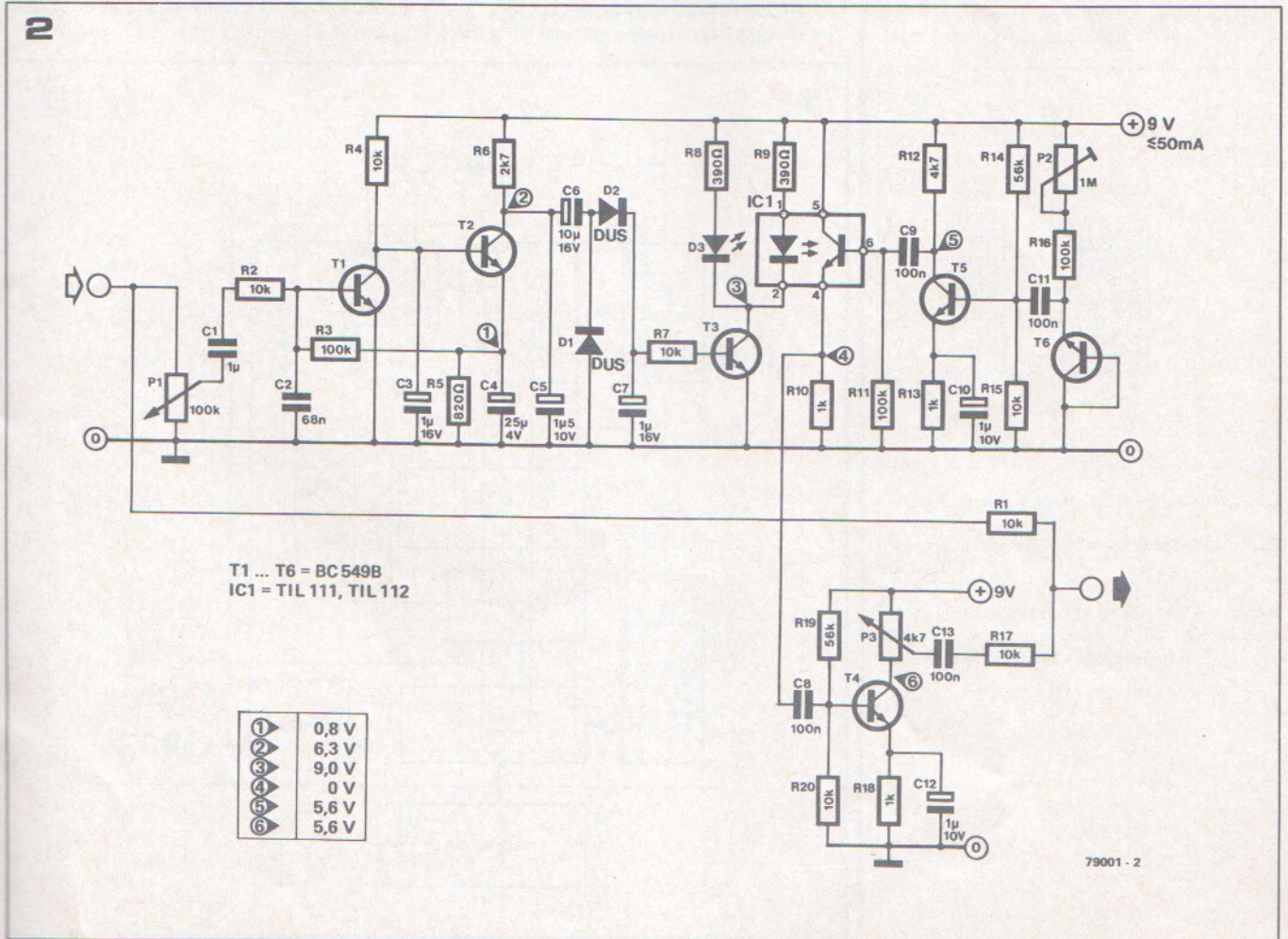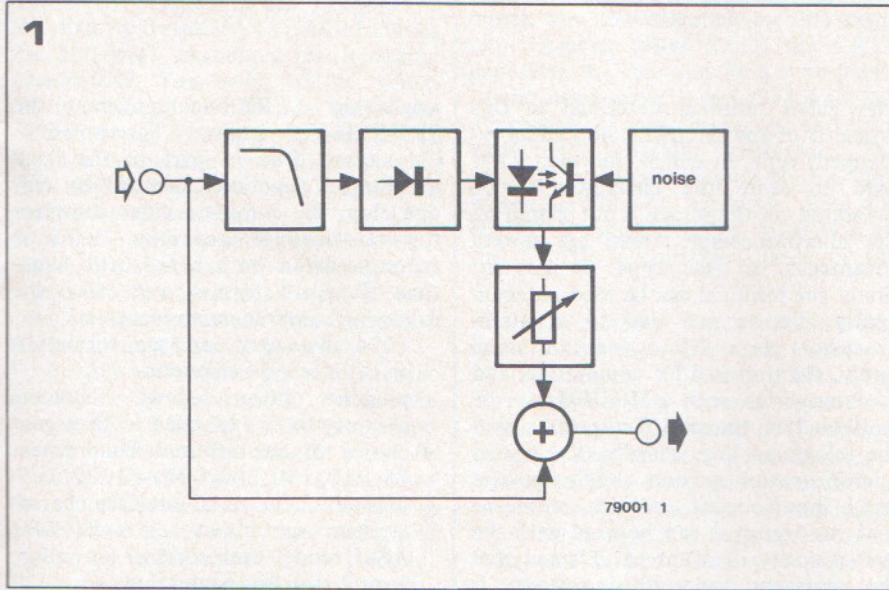
elekterminal

Low cost video terminal for μP/TV typewriter applications

There is no doubt that by far the most convenient and elegant method of displaying data from a microprocessor is on a Visual Display Unit (VDU). When used in conjunction with an ASCII keyboard (such as that described in last month's issue of Elektor) the video interface circuit described here forms a complete video data terminal which can be used with the Elektor SC/MP system or any microprocessor system possessing a serial input/output facility.

The video terminal described in this article is of the serial (i.e. non memory-mapped) type, in which the video RAM used to store the characters to be generated on the screen is not shared by the microprocessor. There are several advantages of this type of system: firstly the terminal can be used independently (i.e. is not tied to a microprocessor) as a 'TV typewriter'. Secondly, the unit is TTY compatible, and in conjunction with a MODEM, can be employed to transmit/receive data over the telephone line. Thirdly, since most microprocessor systems already possess serial input/output routines, it means that the terminal can be used with the vast majority of different μP's and that the necessary device driver software is for the most part already present.

The Elekterminal uses one of the new single-chip CRT controllers, the SF.F 96364 from Thomson-CSF (Sescosem). Due in part to the large number of functions assumed by this one chip, the complete video interface for the terminal uses only 21 ICs, is accommodated on a board little larger than Eurocard format, yet offers the following comprehensive features:

- 1024 characters per page, formatted as 16 lines x 64 characters
- plug-in option allows character memory to be expanded to 16 pages
- choice of six different Baud rates: 75, 110, 150, 300, 600 or 1200
- programmable serial interface characteristics: i.e. choice of 6- or 7-bit ASCII code, even, odd or no-parity, 1- or 2 stop bits generation.
- choice of TTL or RS232C voltage levels

1
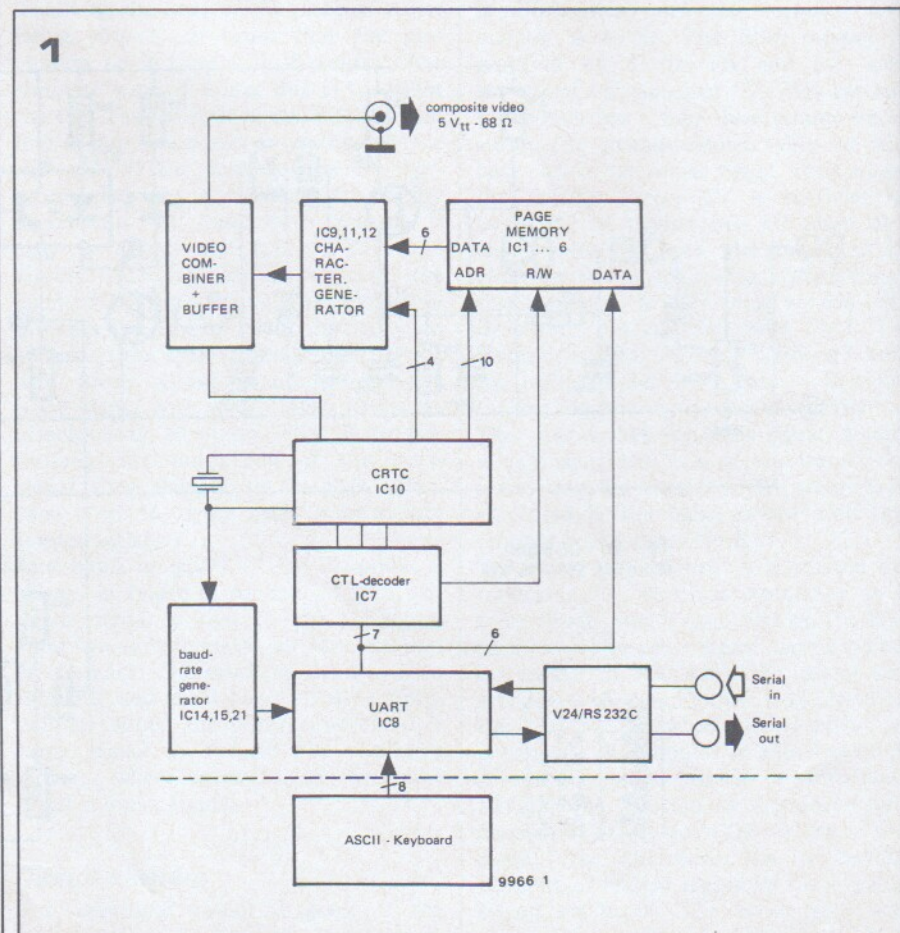
- normal (white on grey) or inverted (black on grey) video signal
- sophisticated cursor control and screen scrolling functions provided in hardware

Block diagram

The block diagram of the Elekterminal is shown in figure 1. A detailed description of the ASCII keyboard was contained in last month's issue, thus the remainder of this article is devoted to the video interface card proper.

The ASCII output of the keyboard is fed directly to the UART. UART stands for Universal Asynchronous Receiver/Transmitter. This is an LSI IC which accepts data, whether in serial or parallel form, from a peripheral device (keyboard, modem) and transmits this data with appropriate serial-parallel or parallel-serial conversion to the CPU or video interface. Basically the UART allows the keyboard, VDU and CPU to communicate with one another. A more detailed description of this important IC is contained later in the article.

To be able to transmit data at different speeds, a programmable Baud rate generator is required. As was explained in the article on the cassette interface (Elektor 36), the Baud rate is defined as the total number of bits including control bits such as stop and parity bits which are transmitted in one second. The programmable Baud rate generator generates a number of frequencies which are 16 x the desired Baud rate.

The output frequencies are derived by dividing down a clock input obtained from the crystal oscillator of the CRTC. The CRTC is without doubt the most important component in the entire circuit. CRTC stands for Cathode Ray Tube Controller, however it might be more accurate to describe the chip as a 'video processor'. However one calls it, the device is another LSI IC which performs a wide variety of control functions with a minimum of peripheral hardware. In the past, video interface cards required a veritable mountain of discrete logic ICs to perform the tasks which are now assumed by this single chip. Among other things the CRTC generates the line and field sync pulses of the video signal, is responsible for the addressing of page memory, and controls the character generator. The chip also provides cursor control and screen scrolling in hardware. Like the UART, a more detailed discussion of the CRTC is contained later in the article.

The page memory, which holds the data to be displayed on the screen, is formed by a number of static RAMs. The entire memory is scanned once every frame (20 ms). The ASCII code which is stored in page memory is converted into a form suitable for display with the aid of a character generator; after parallel-serial conversion it is then added to the horizontal and vertical sync pulses in the video combiner. The latter provides a 5 V_{pp} signal which can be fed direct to a video monitor or else via a

VHF/UHF modulator (such as that contained in the October issue of Elektor this year) to the aerial input of a domestic TV receiver.

The only section of the circuit which remains to be discussed is the CTL decoder. This is basically a ROM which decodes the ASCII character transmitted by the UART and informs the CRTC whether it is a control signal (non-printing code) or a character to be displayed on the screen.

UART, character generator and CRTC

Since many readers may be unfamiliar with these important devices, it is worth while taking a closer look at just exactly how they operate.

UART

The block diagram of the AY-5-1013 UART which is used in the Elekterminal is shown in figure 2. In fact the UART can be thought of as two ICs (a transmitter and a receiver) which are housed in the same package and which combine certain functions in order to economise on the number of terminal pins. The UART is basically a device which provides asynchronous control of data communications, that is to say it is capable of both receiving and transmitting data at different rates, as well as performing parallel-serial and serial-parallel conversions, adding or deleting the necessary control and error detecting bits as required.
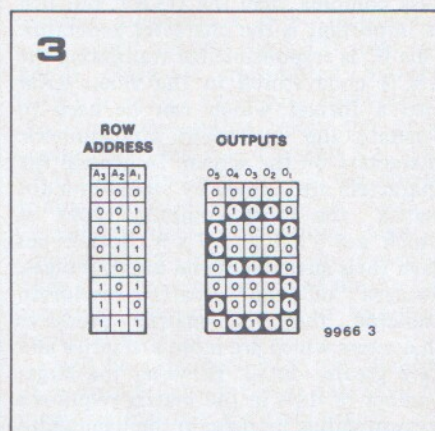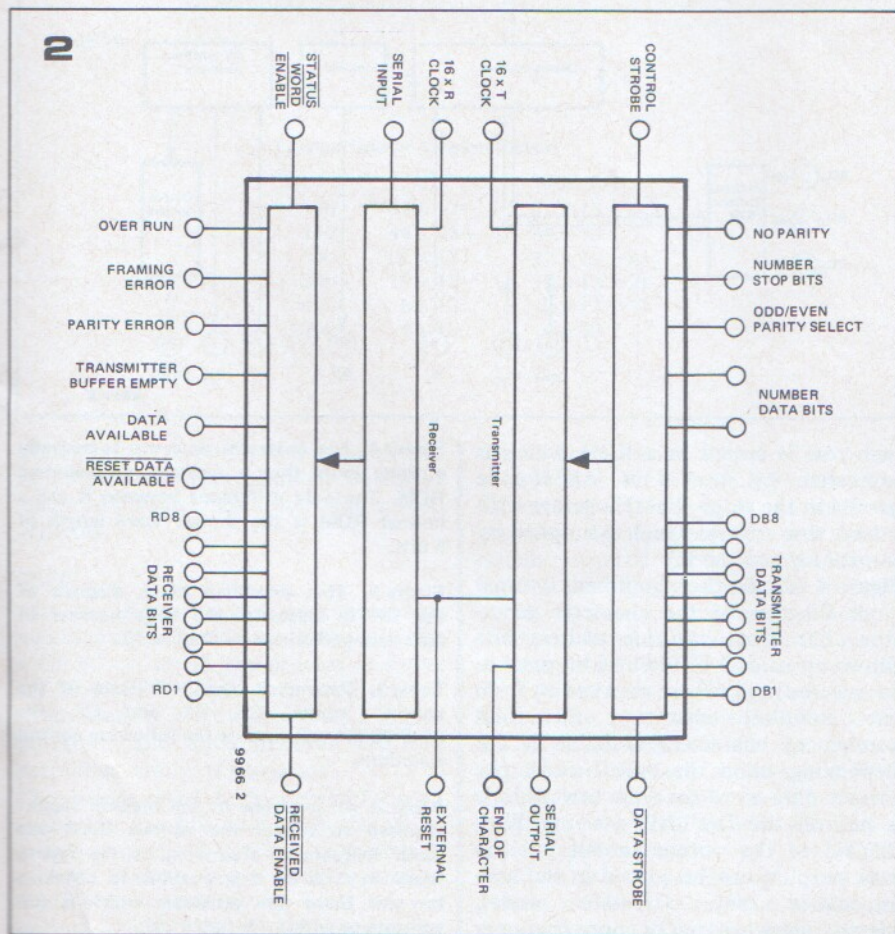

Figure 1. Block diagram of the Elekterminal.

Figure 2. Simplified internal block diagram of the UART.

Figure 3. Characters are generated by means of an 8 x 5 dot matrix. The top row of dots remains permanently extinguished.

The transmitter and receiver have independent reference clock inputs, which determine the rate at which the data flow occurs, the UART can thus be used for both code- and speed conversion. The Baud rates are determined by the output frequency of a programmable Baud rate generator contained on-chip. The output frequency of the Baud Generator is 16 x the Baud rate.

Data are fed to the transmitter section of the UART (e.g. the ASCII output of the keyboard) in parallel form. The UART converts the parallel code into a serial data stream, adding the necessary start-, stop- and (if desired) parity bits. The system user can program the format of the transmitted or received serial character to suit his own requirements. That is to say that he has the choice of one or two stop bits, odd, even or no parity bit, and of selecting a 5, 6, 7 or 8-bit data word. The receiver section of the IC does exactly the opposite of the transmitter, i.e. it deletes the start and stop bits from the received serial character, checks for parity errors (which are flagged by setting the parity-error output), and presents the data in parallel form at the data outputs. When using the UART as a Baud rate- or code converter the data at the receiver outputs are fed to the data inputs of the transmitter; code conversion is performed by means of a ROM decoder connected between the two sections.

Character generator

Less complex than the UART, but just as important is the character generator. This IC is responsible for translating the ASCII code stored in the video RAM into a format which can be used to generate the equivalent alphanumeric characters on the screen. In general the characters are formed by means of a dot matrix, the most common types of which are 5 x 7 and 7 x 9. Both types have their advantages and disadvantages. Because of the greater resolution available the 7 x 9 matrix produces characters which are more attractive and have greater detail. However the larger number of dots in the matrix requires a corresponding increase in the bandwidth of the video signal. With 64 characters per line this bandwidth is several Megahertz too large for conventional TV receivers, and results in poor picture definition. For this reason the 7 x 9 matrix is generally reserved for use with video monitors.

Although the characters generated by a 5 x 7 matrix are somewhat simpler, it is still possible to obtain excellent definition using a normal domestic TV receiver which has a video input. Even with the unavoidable picture degradation caused by a VHF/UHF modulator, the legibility of the resulting display is still perfectly satisfactory.

The format of the matrix produced by the character generator is illustrated in figure 3. The information contained in
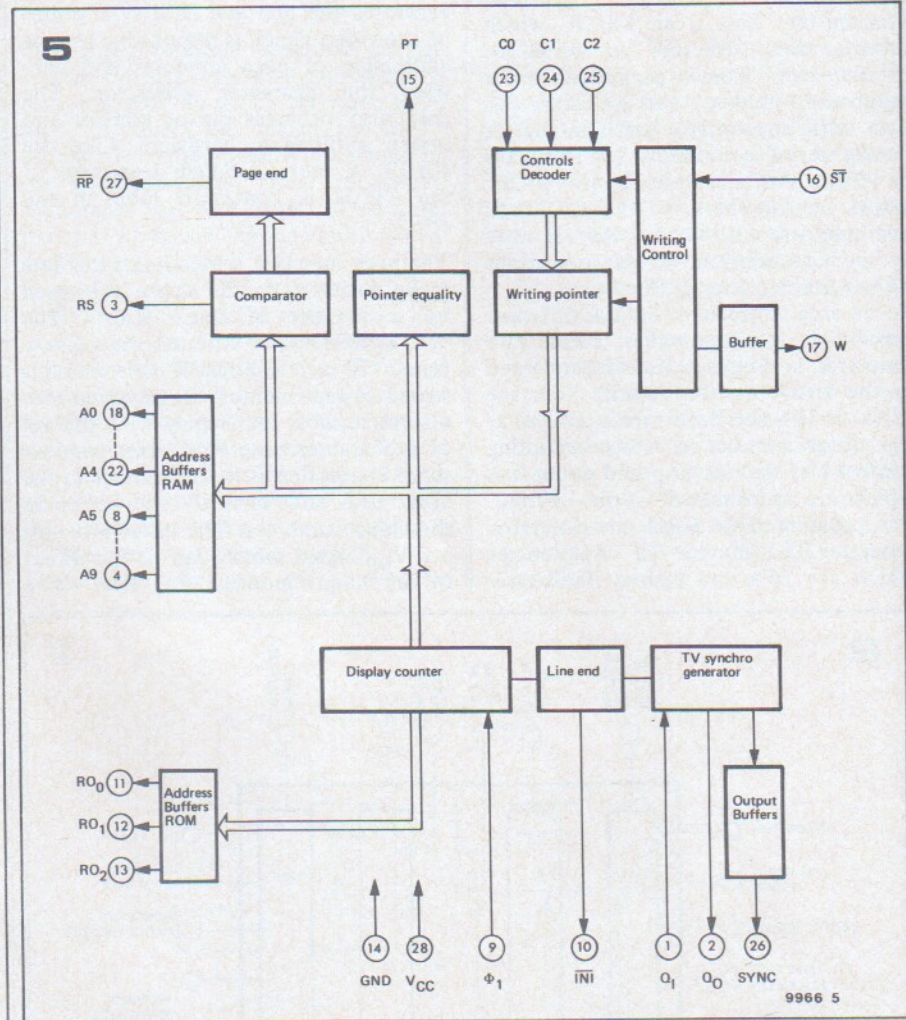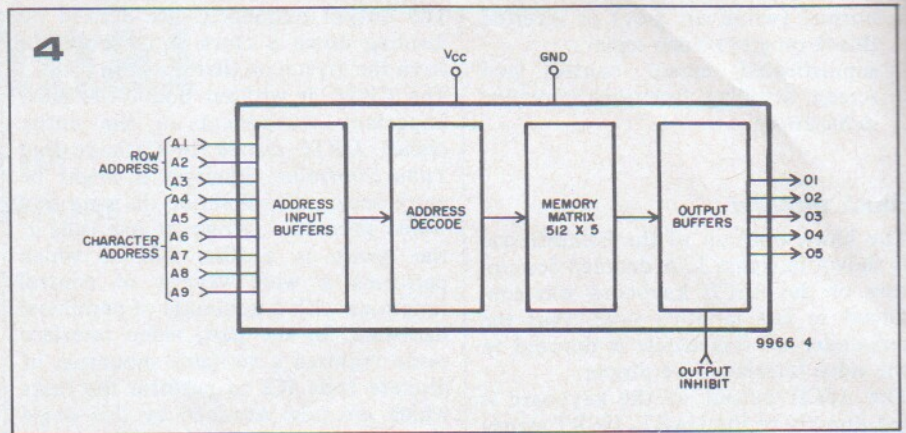
each row is stored in a ROM which is addressed by the 6-bit ASCII code (stored in the video RAM) together with a 3-bit row address which is supplied by control logic in the IC.

Figure 4 shows the simplified internal block diagram of the character generator. The total of nine address bits allows up to 512 different 5-bit rows to be selected; since 8 are required to form one complete character, the total number of characters available is 64. Depending upon the ASCII code, the correct data word for each row address is put on the five data outputs. With the aid of the output inhibit pin the data outputs can be placed in the high impedance state (tri-state mode), thereby allowing two or more character

Figure 4. The character generator is basically nothing more than a specially programmed ROM. The only difference between it and a normal ROM is the shorter word length of 5 bits.

Figure 5. This simplified block diagram of the CRTC illustrates the large number of functions performed by the one IC.

Table 1. Depending upon the state of the control inputs, C0, C1 and C2, the SF.F 96364 will execute the following control functions.

Table 2. An overview of the division factors required to obtain the various Baud rate clock frequencies. Rounding up the figures listed in table 2a it is possible to obtain a low-cost Baud rate generator which is still accurate to within 1% (table 2b).

Table 1.

	C_2 C_1 C_0	Execution time ms
Page erase and cursor home (top-left)	0 0 0	132
End of line erase and cursor return (at left)	0 0 1	8.3
Line feed (cursor down)	0 1 0	8.3
Inhibition of the character sent	0 1 1	8.3
Cursor left (one position)	1 0 0	8.3
Erasure of cursor-line	1 0 1	8.3
Cursor up (one position)	1 1 0	8.3
Normal character	1 1 1	8.3

Table 2a

Baud rate	f UART	division factor 1000 MHz	division factor 1008 MHz
75	1200 Hz	833.33	840
110	1760 Hz	568.18	572.73
150	2400 Hz	416.67	420
300	4800 Hz	208.33	210
600	9600 Hz	104.17	105
1200	19200 Hz	52.08	52.50

Table 2b

Comparison of division factors

Baud rate	1 MHz	1008 MHz
75	64 x 13	64 x 13 (+8)
110	44 x 13 (−4)	44 x 13
150	32 x 13	32 x 13 (+4)
300	16 x 13	16 x 13 (+2)
600	8 x 13	8 x 13 (+1)
1200	4 x 13	4 x 13

generators to be connected in parallel so as to provide the remaining 64 ASCII characters, lower case letters and special symbols.

CRTC

Almost all the major microprocessor manufacturers have already brought out a CRTC or are in the process of doing so. Most CRT Controllers are designed to be used in conjunction with a microprocessor, and some are even tied to a particular family of processor.

The device used here, however, is an exception to this rule, and the video interface card, of which it is the heart, can be used to form an independent VDU/TV typewriter as well as an output terminal for any microprocessor

which has a serial output. The device in question is the SF.F 96364 from Thomson-CSF, which, as we shall see, provides all the control and timing signals required for screen display, as well as providing a number of sophisticated screen management facilities (cursor control and screen scrolling etc.) A simplified block diagram of the SF.F 96364 is shown in figure 5.

One of the most important tasks of the CRTC is to generate the sync pulses required to display a video signal. With the aid of a simple on-chip crystal oscillator the SF.F 96364 provides a close approximation to the CCIR standard sync signal. Line and field sync pulses are both combined in the one sync waveform. The sync generator also

drives the display counter which is responsible for the addressing of the character generator (i.e. providing the correct row address) and of the video RAM (page memory). In addition, the display counter provides information for the cursor- and page-end comparators. The cursor comparator supplies a signal which ensures that the cursor appears at the correct point on the screen. The page-end comparator allows the amount of addressable video RAM to be extended to include extra pages, since the RS output is used to enable the VDU to 'turn a page' in mid-screen. The RP output clocks the counter used to address the additional pages of character memory.

A detailed explanation of how the video RAM is extended will be contained in a subsequent article describing an add-on circuit which will permit the use of up to 16 pages of memory.

The above mentioned functions of the CRTC are of course indispensable, however the most important features of the device are almost certainly the screen formatting control functions which are available in hardware. Many less sophisticated video interface cards rely upon software routines to provide control of cursor and screen scrolling, which means that they are necessarily tied to a microprocessor. The SF.F 96364, however, can provide fairly sophisticated screen control options on-chip, allowing it to function independently.

As was stated in the description of the block diagram in figure 1, the CTL decoder provides a 3-bit instruction code which informs the CRTC that the ASCII character transmitted by the UART is in fact a non-printable control character. Depending upon the code applied to pins C_0, C_1 and C_2, the CRTC will perform one of the cursor control functions listed in table 1. Certain control instructions take a comparatively long time to execute, since they have to be carried out during blanking intervals so as to minimise display distortion.

As will be explained later, the number of control functions can be extended by manipulating the W-(Write-)signal; this facility is exploited in the Elekterminal.

Circuit

The 21 ICs and associated components which are shown in figure 6 form the complete circuit of the Elekterminal. All that is required to render the terminal operational is the addition of an ASCII encoded keyboard. Page memory, which holds the ASCII version of the characters to be displayed on the screen, is 6 bits wide and is formed by six 2102A4 1 K x 1 RAMs. The '4' in the type number indicates the access time of the device, which in this case is 450 ns. If character memory is to be expanded to several pages, it may well be worth investing in low power
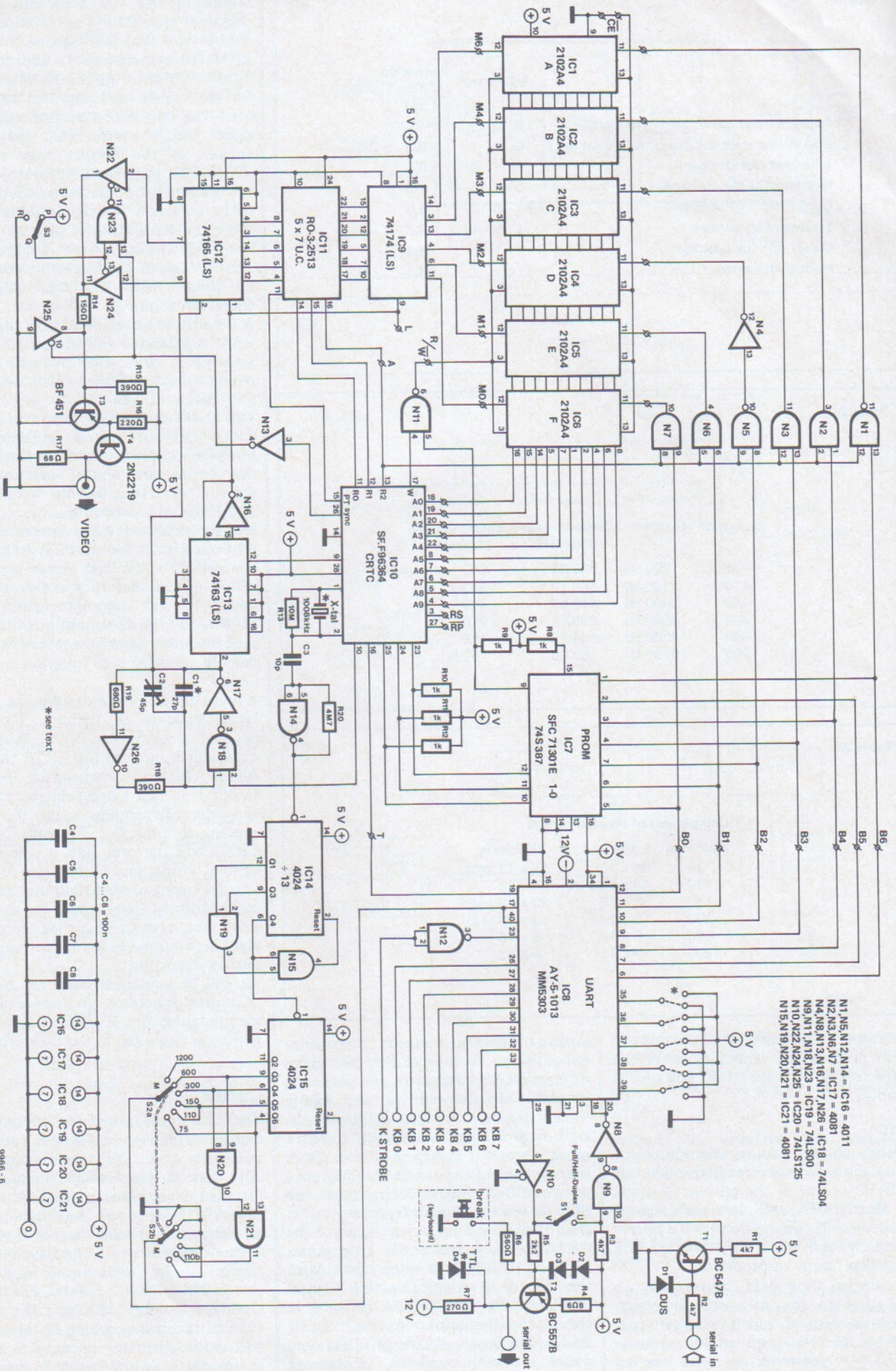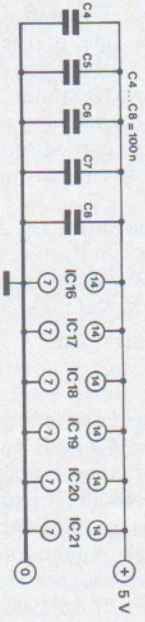
6

Table 3.

PIN	LEVEL	TRANSMITTED OR RECEIVED FORMAT
35	1	No parity bit
35	0	Transmitted parity bit
39	1	Even parity
39	0	Odd parity
36	1	2 stop bits
36	0	1 stop bit
37 38	0 0	5 bits/character
37 38	0 1	6 bits/character
37 38	1 0	7 bits/character
37 38	1 1	8 bits/character

Table 4.

Function	Key	corresponds to
line-feed	LF	CTL J
carriage-return + erase to end of line	CR	CTL M
cursor up	VT	CTL K
cursor down	LF	CTL J
cursor left	BS	CTL H
cursor right	HT	CTL I
home cursor	FS	——
home cursor + page erase	FF	CTL L
scroll up (cursor down)	ESC	CTL [
carriage return (no erasure)	– – (GS)	CTL]
erase current line	– – (SUB)	CTL Z

Table 5.

Positive logic

Address	O_3	O_2	O_1	O_0
0 to 127	1	0	0	0
128 to 135	0	0	1	1
136	0	1	0	0
137	0	1	1	1
138	1	0	1	0
139	0	1	1	0
140	1	0	1	1
141	1	0	0	1
142 to 153	0	0	1	1
154	1	1	0	1
155	0	0	1	0
156	0	0	0	0
157	0	0	0	1
158, 159	0	0	1	1
160 to 254	1	1	1	1
255	0	0	1	1

memories (2102AL4), since they can lead to a saving in current consumption of up to roughly 30%.

Since each character is formed by 8 rows of 5 bits, the ASCII code stored in the page memory is read out 8 x every frame. With 64 characters per line, the memory is scanned in blocks of 64 words. IC10, i.e. the CRTC, ensures that the same block is scanned 8 times in succession, and that the character generator is provided with the correct row addresses. The outputs of the memory are not connected directly to the character generator, but to an intermediate latch (IC9). The memory address can thus remain one step ahead of the position on-screen, which means it has ample time to set up the following ASCII code on its outputs.

The 5-bit parallel 'row' data from the character generator is fed to a shift register (IC12), where it is converted into serial form and thus becomes suitable for video display. This shift register is driven by a 'dot-clock' with a frequency of approx. 11 MHz. The dot-clock generator is formed by N17, N18 and N26. Since all eight rows of a character must be positioned directly beneath one another, the dot-clock generator is synchronised by the CRTC. This is achieved with the aid of the INI line (see figure 5) which goes low after the 64th character, thereby stopping the dot-clock generator until the following line sync pulse.

All memory addressing is clocked by the dot-clock, since the 'character-clock', which, via the Φ1-input, drives the address counter in the CRTC, is derived from the dot-clock signal via a divide-by-eight counter (IC13).

The frequency of the dot-clock, which can be varied by means of C2, determines the width of a character: the lower the frequency the wider the character. The minimum usable frequency is determined by the available space on the TV screen. If too low a fre-quency is chosen, the characters will run off the edge of the screen. On the other hand, if the frequency is too high, the characters will be compressed into one portion of the screen, with a consequent loss of definition and legibility. Thus it is important that C2 be adjusted so as to obtain an optimal picture on-screen.

The spacing between character lines is regulated by the CRTC, which blanks the video signal for the period of four line scans. Between successive character lines there is therefore a space of half a line. Spacing between the actual characters is provided by the shift register, IC12. Since this is an eight-bit shift register and the character generator output is only five bits wide, each character can be preceded by two unmodulated dot columns and followed by one such column. Between each character there is thus a space three dots wide. The fact that, including spaces, each character is in fact eight bits wide explains why the character clock is derived from the dot-clock via a divide-by-eight counter.

The serial data stream output by the shift register is also available in its inverted form. One can therefore choose between a positive (white on grey) or negative (black on grey) video signal.

The video combiner is built round N22 . . . N25. Depending upon the position of S3, N22 . . . N24 provide a video signal of appropriate polarity. N23 not only inverts the signal at S3, but also inhibits the inverted video signal during the sync period. This causes the current through N25 during the sync pulses to be limited to an acceptable value.

The voltage divider network formed by R14 and R15 determines the ratio between the amplitude of the video signal and that of the sync signals. With the values shown the black level is around 30%.

Figure 6. Complete circuit diagram of the video interface for the Elekterminal. With the addition of an ASCII keyboard, which is connected to the K strobe- and KB0 . . . KB7 lines, the circuit forms a complete video data terminal.

Table 3. The programmable serial interface characteristics of the UART. This table applies both to the AY-5-1013 and to the MM5303. A recommended bit format is shown shaded; this format corresponds to the arrangement of wire links illustrated in figures 6 and 8.

Table 4. In addition to the control functions listed in table 1, the 4-bit PROM decoder offers several extra possibilities. These extra control functions can be generated either by special keys or with the aid of the control key and one of the data keys.

**Table 5.** The program for the PROM decoder IC7.

The video combiner is followed by a buffer stage which has an output impedance of 68 Ω, and to which one can directly connect a length of coaxial cable. Assuming that the cable is terminated with the correct impedance it is possible to use lengths of over 10 m without any problems. The buffer stage has the effect of increasing the black level to around 35%; this can of course be corrected by altering the values of R14 and R15, however the increase in sync signal level does no harm and the adjustment is not really worth the bother.

So much for the circuitry which is responsible for actually generating the video signals; there remains the circuit which allows the unit to communicate with peripheral devices such as the CPU and/or keyboard. The most important interface element is of course the UART, the basic operation of which has already been described. As was stated then, the rate at which data is transmitted by the UART is determined by a clock signal, whose frequency is 16 x the desired Baud rate. Normally a monolithic Baud rate generator is employed to provide the clock signal, however these ICs are still fairly expensive and require a 1 MHz crystal to supply the basic clock frequency from which the 'x 16' clock signals are obtained by means of frequency division.

An obvious alternative is to make use of the crystal oscillator for the CRTC to provide the necessary clock frequencies. This can be done quite simply by first of all amplifying the signal at the output of the oscillator (with the aid of N14), and then feeding it to a programmable divider (IC14 and IC15). The accuracy of the clock frequencies obtained thereby is better than 1%. Generally speaking, the methods adopted for the transmission of data allow a reasonable tolerance in the accuracy of signals, so that a deviation in the region of 1% remains a perfectly acceptable figure. Table 2 lists the relationship between the UART frequencies and the (theoretical) division factors (2a) obtained with crystal frequencies of 1 MHz and 1008 kHz. The manufacturer of the SF.F 96364 recommends a crystal frequency of 1008 kHz in order to avoid interference from the mains frequency. In practice, however, a 1 MHz crystal is perfectly satisfactory.

As can be seen from table 2, keeping to an accuracy of 1% and rounding up to the nearest whole figure produces identical division ratios for both crystal frequencies. The result is a Baud rate generator at roughly 20% the price of a monolithic type. Results obtained in practice have failed to indicate that this approach has any untoward effect upon the performance of the circuit.

Once provided with suitable clock frequencies, the UART can receive and transmit data at any one of six different
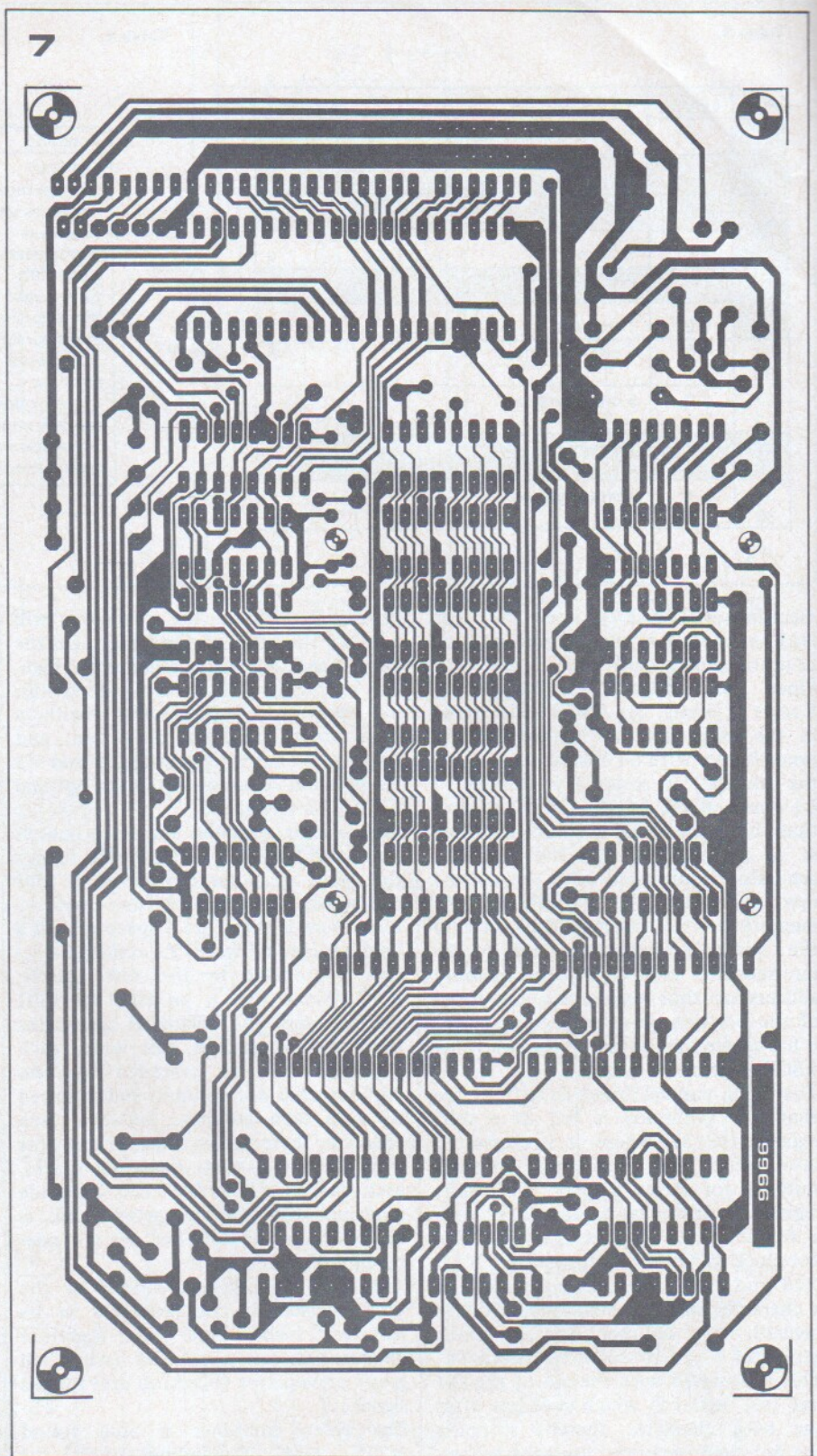
Baud rates, which are selected by means of switch S2. An important point about most UARTs is that the output logic levels are often not TTL-compatible. The most common voltage levels used are the so-called RS232C and V24 norms. These two norms, which are virtually identical and more or less interchangeable, have the advantage of minimum signal levels of +5 V (logic '1') and −5 V (logic '0') and maximum levels of + and −15 V respectively. Thus they clearly have a much greater noise immunity than TTL logic levels.

The circuit described here attempts to reach a compromise solution by employing a discrete interface design which is compatible with both RS232C/V24 and TTL voltage levels. If the output signal is required to drive TTL loads, then D4 should be included. This diode limits the output voltage to −0.6 V. Without the diode the output voltage can swing between +5 V and −12 V. The output impedance is deliberately held low in order to facilitate matching with the cable.

As was mentioned earlier, the format of

8
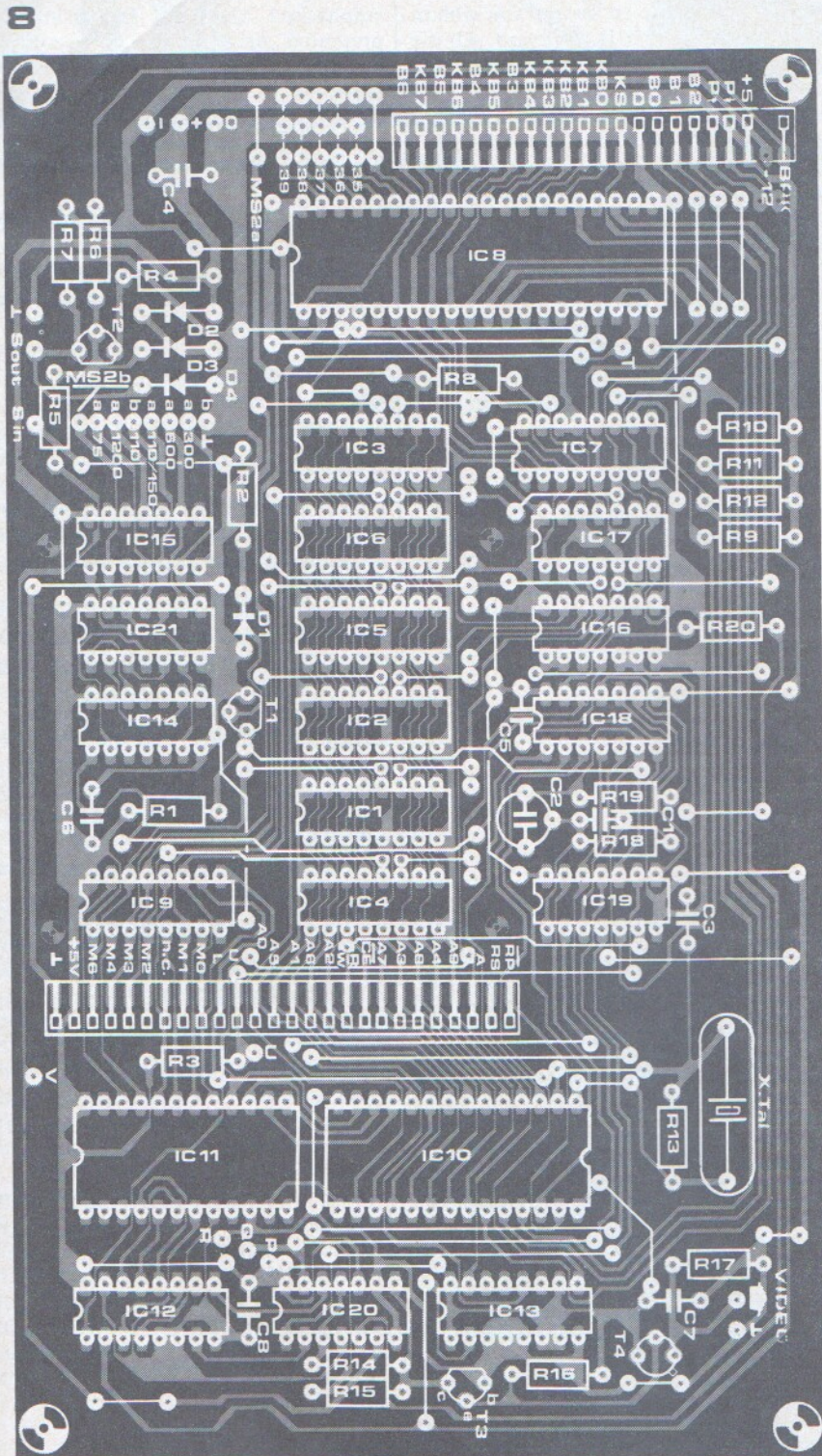
Parts list to figure 6:

Resistors:
R1 . . . R3 = 4k7
R4 = 6Ω8
R5 = 2k2
R6 = 560 Ω
R7 = 270 Ω
R8 . . . R12 = 1 k
R13 = 10 M
R14 = 150 Ω
R15 = 390 Ω
R16 = 220 Ω
R17 = 68 Ω
R18 = 390 Ω
R19 = 680 Ω
R20 = 4M7

Capacitors:
C1 = 27 pF (see text)
C2 = 45 pF trimmer
C3 = 10 pF
C4 . . . C8 = 100 n

Semiconductors:
D1 . . . D4 = DUS
T1 = BC 107B, BC 547B or equ.
T2 = BC 177B, BC 557B or equ.
T3 = BF 451
T4 = 2N2219

ICs:
IC1 . . . IC6 = 2102-1,2102A4,
  2102AL4
IC7 = SFC 71301E 1-0 (pre-
  programmed) or equivalent
  eg 74S387 (to be programmed
  as shown in table 5)
IC8 = AY-5-1013,MM5303
IC9 = 74LS174
IC10 = SF.F 96364 (Sescosem)
IC11 = RO-3-2513
IC12 = 74LS165
IC13 = 74LS163
IC14,IC15 = 4024
IC16 = 4011
IC17,IC21 = 4081
IC18 = 74LS04
IC19 = 74LS00
IC20 = 74LS125

Miscellaneous:
S1 = SPST switch
S2 = 2 pole 6 way switch
S3 = SP changeover switch
P.c.b. connectors (female): type
  ITT-Cannon G09
1 x 22 way (keyboard)
1 x 26 way (extension card)
For keyboard cable (male): type
  ITT-Cannon G09
1 x 22 way
X1 = 1008 kHz or
  1000 kHz crystal

the serial output/input signal can be programmed by the system user. The number of stop/start bits, choice of parity bit and of data word length can be selected by making the appropriate connections to pins 35 . . . 39. The details are listed in table 3, where a preferred format (7-bit code with even parity) is shown shaded. If desired the parity bit can be omitted (no parity) since although the UART checks for parity errors in the received signal, the parity-error output is not brought out externally. Thus the parity bit is only of use to the device receiving a character *transmitted* by the UART.

The Elekterminal is capable of being operated in both the full-duplex and half-duplex mode. In a full-duplex system, where the terminal is linked to a µP, the CPU and terminal communicate in both directions simultaneously. That is to say that the computer is programmed to echo what is transmitted (from the keyboard via the UART) back to the terminal display. In half-duplex systems the terminal is normally wired up so that the screen

responds directly to the keyboard. Switching between half- and full-duplex is accomplished with the aid of S1, which is included between the serial output- and input pins of the UART.

The UART relays the ASCII code from the keyboard (or CPU) onto the data bus lines B0 . . . B6, where it is picked off by the CRTC and character memory. Before the data reaches the RAMs, however, it is converted from 7- to 6-bit ASCII code; bit 5 is ignored and bit 6 inverted. In addition, gates N1 . . . N7 offer the possibility of forcing the 'space' code (100000) onto the data inputs of the RAMs, so that, if the appropriate control code is applied to the CRTC, an entire line or the complete screen can be erased. These are in fact only two of the many control functions which the Elekterminal possesses.

The 7-bit ASCII control codes are detected and decoded by a 256 x 4 ROM (IC7), which forms the CTL decoder of figure 1. The ASCII code is placed on the address inputs of this ROM and the code which appears at the data outputs is fed to the C0, C1 and C2 control lines of the CRTC.

Table 1 has already listed a number of the control functions offered by the CRTC. However by utilising the read/write line of the RAMs it is possible to extend these. Table 4 provides an overview of all the various cursor control and screen scrolling functions which are provided by the Elekterminal. The majority of these functions can be selected by an individual key on the Elektor ASCII keyboard which was published last month. However, with the exception of 'home cursor', all the above functions can also be obtained using the control key and the appropriate data key, which means that the video interface is compatible with keyboards other than the Elektor model.

The PROM decoder for the CRTC is programmed as shown in figure 5. This device is available from a number of manufucaturers under the type number 74S387. Since only 128 combinations are possible with a 7-bit code, only half of the PROM is used.

Printed circuit board

The printed circuit board for the video interface card (see figures 7 and 8) is not much larger than Eurocard format, but nonetheless is single-sided. Because of this there are a considerable number of through connections to be made (approx. 60 in all), however the extra effort required is more than compensated for by the lower cost of a single-sided board. The p.c.b. has been specially designed to accomodate an extension board to increase the number of memory pages to 16. The latter (which will appear shortly) simply plugs into the video interface card with the aid of edge connectors. At the left hand

end of the video interface board is a simple connector to accept the ribbon cable from the ASCII keyboard. All the connections to the keyboard, i.e. including supply lines, can be made via the connector socket. This connector is also used to bring out the data lines from the UART. These connections will be required when incorporating the extension board.

The second p.c.b. connector provides access to all the address and data lines of the character memory and to two lines which are used to enable the RAMs to be addressed properly. This connector is designed to accomodate the

memory extension board. It should be emphasised that the Elekterminal as presented here represents a complete output peripheral, which can be extended to accomodate more memory by plugging in one or more additional cards. The only on-board modification required is the removal of one through connection.

The UART (IC8) is programmed by means of wire links. The connections indicated on the component overlay correspond to the recommended format listed in table 3.

Connection to a TV

Not every TV has an input for an unmodulated video signal, but in those cases where one is present, or if a video monitor is being used, the amplitude of the terminal output signal has to be adjusted to suit the sensitivity of the input in question. This is best done as follows:

The video signal is fed to the TV receiver or monitor via a length of coaxial cable (50 . . . 75 Ω). The 'receiver-end' of the cable should terminate in a low impedance. A 100 Ω potentiometer is ideally suited for this purpose. The potentiometer can then be used to adjust the signal amplitude to a suitable value. Naturally, an alternative solution will have to be found in the case of sets which are provided with an internal terminal impedance.

If an input for unmodulated video signals is not available the output must be fed to a VHF/UHF modulator such as that published in the October issue of Elektor this year. Due to the large bandwidth of the video signal a certain degradation of picture quality is inevitable, however the resultant definition is still quite acceptable for the type of application for which the Elekterminal is intended. With or without a modulator, the signal amplitude should be adjusted so that the picture is 'sync-ing' with both positive and negative polarity video signals. This can easily be checked by changing the polarity several times in succession. One should also first ascertain that the line oscillator of the TV receiver is correctly tuned.

Supply

Using normal memory ICs the current consumption of the circuit is around 750 mA (5 V). If low power memories are used, however, consumption drops to around 550 mA (5 V). The −12 V supply draws well under 100 mA. This means that the circuit could be powered using the spare capacity of an SC/MP system, assuming the latter was not driving a large amount of additional memory. Alternatively, a better idea might be to use the SC/MP power supply design to form a separate supply for the Elekterminal, especially if one bears in mind that extending the number of character memory pages will push up the current consumption. ◀
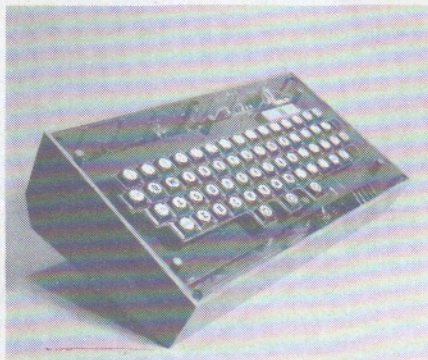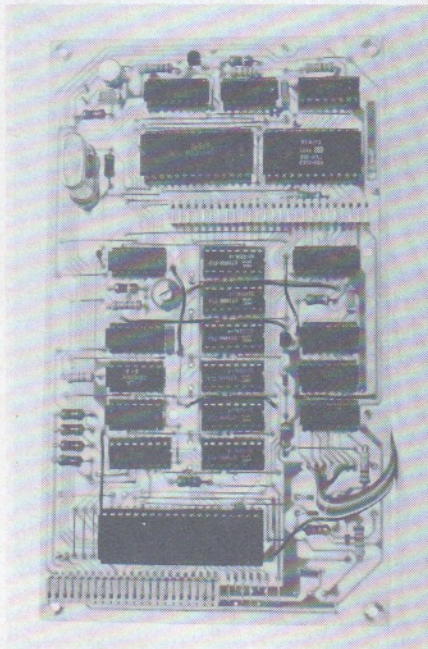
RAM diagnostic

An SC/MP program to test large sections of RAM

ess
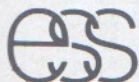
The more RAM which one adds to a μP system, the more difficult it becomes to trace any faults which may occur on one of the memory boards. Think, for example, of how tedious it would be to have to step through every location of a 4 K RAM card! Fortunately, however, with the aid of the right software, we can make the microprocessor itself do all the hard work. After all, computers are supposed to be ideally suited to taking over boring and repetitive tasks from humans. The following program can be used to test any block of memory larger than ½ K and will indicate the location of any faults which exist.

H. Huschitt

A static Random Access Memory basically consists of a large number of flip-flops and address decoders. These decoders are either integrated on the RAM chips themselves, or else are formed by discrete (TTL and CMOS) logic ICs. However, if somewhere within a section of memory there is a faulty IC, a short circuit or a loose contact, there is a good chance that the program which is contained in that section of memory will fail to function properly. A suitable debug program must therefore be able to check for such a malfunction.

The flow diagram of the RAM-diagnostic program is shown in figure 1.
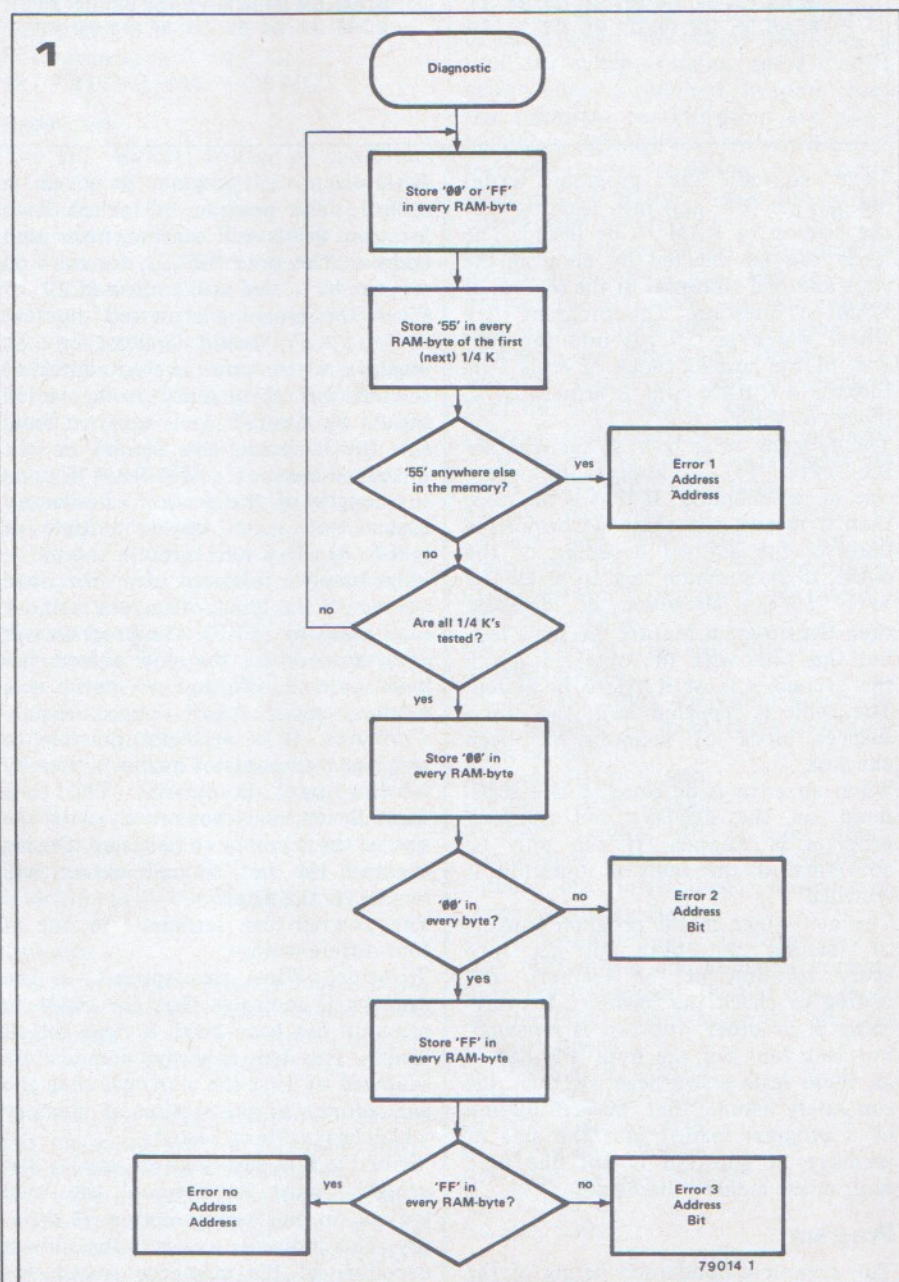
Figure 1. Flow diagram for the RAM-diagnostic program, showing the order of the different tests.

Table 1. For reasons of space, the program is listed in condensed form. The complete program is also recorded on the Elektor Software Service record ESS-001.

2

```
            RAM — DG

0C00 C4 5E C9 06 C4 3D C9 05
0C08 C4 00 C9 FF C9 00 C4 80
0C10 C9 04 C9 03 C9 02 C9 01
0C18 C4 3E CA 1D C4 00 37 C4
0C20 55 33 3F C2 01 CA 0A CA
0C28 10 C2 02 CA 0B CA 11 3F
0C30 C2 01 CA 13 C2 02 CA 14
0C38 C4 00 CA 12 C2 10 CA 0E
0C40 C2 11 CA 0F C4 0D 37 C4
0C48 B4 33 3F C4 55 CA 12 C2
0C50 0A CA 0E C2 0B CA 0F C2
0C58 01 CA 13 C2 0B CA 14 C4
0C60 0D 37 C4 B4 33 3F C2 0F
0C68 E2 01 9C 08 C2 0F E2 02
0C70 9C 02 90 22 C2 01 CA 13
0C78 C2 02 CA 14 02 C2 0E F4
0C80 01 CA 0A CA 0C C2 0F F4
0C88 00 CA 0B CA 0D C4 0D 37
0C90 C4 DC 33 3F 90 A2 C4 00
0C98 CA 12 C2 10 CA 0E C2 11
0CA0 CA 0F C4 0D 37 C4 B4 33
0CA8 3F C2 10 CA 0C C2 11 CA

0CB0 0D C4 0D 37 C4 DC 33 3F
0CB8 C4 FF CA 12 C2 10 CA 0E
0CC0 C2 11 CA 0F C4 0D 37 C4
0CC8 B4 33 3F C2 10 CA 0C C2
0CD0 11 CA 0D C4 0D 37 C4 DC
0CD8 33 3F C2 01 CA 13 C2 02
0CE0 CA 14 C4 AA CA 12 90 1B
0CE8 C2 0C CA 10 C2 0D CA 11
0CF0 C2 12 E4 55 9C 0A C4 00
0CF8 CA 13 C2 0F CA 14 90 03
0D00 40 CA 14 C4 01 31 C4 07
0D08 35 C2 12 01 40 E4 AA 9C
0D10 0A C4 54 C9 00 C4 5C C9
0D18 FF 90 24 40 E4 55 9C 0A
0D20 C4 06 C9 00 C4 00 C9 FF
0D28 90 15 40 98 0A C4 4F C9
0D30 00 C4 00 C9 FF 90 08 C4
0D38 5B C9 00 C4 00 C9 FF C4
0D40 79 C9 06 C4 50 C9 05 C9
0D48 04 C9 02 C4 5C C9 03 C4
0D50 00 C9 01 C4 0E 37 C4 1C
0D58 33 3F C4 A0 CA 1D C2 10
0D60 CA 01 C2 11 CA 02 C4 00
0D68 37 C4 55 33 3F C4 00 C9

0D70 FF C9 00 C9 05 C9 06 C4
0D78 0E 37 C4 1C 33 3F C2 13
0D80 CA 01 C2 14 CA 02 C4 00
0D88 37 C4 55 33 3F C4 00 C9
0D90 FF C9 00 40 E4 AA 98 0B
0D98 40 E4 55 98 06 C4 00 C9
0DA0 01 C9 02 8F FF 8F FF C4
0DA8 0E 37 C4 1C 33 3F C4 0D
0DB0 37 C4 02 33 3F C2 0E 31
0DB8 C2 0F 35 C2 12 C9 00 C2
0DC0 0E E2 13 9C 06 C2 0F E2
0DC8 14 98 0F 02 C2 0E F4 01
0DD0 CA 0E C2 0F F4 00 CA 0F
0DD8 90 DB 3F 90 D8 C2 0C 31
0DE0 C2 0D 35 C1 00 E2 12 C9
0DE8 02 01 C2 12 E4 55 98 05
0DF0 40 98 0C 90 03 40 9C 07
0DF8 C4 0C 37 C4 E7 33 3F C2
0E00 0C E2 01 9C 09 C2 0D E2
0E08 02 9C 03 3F 90 CF 02 C2
0E10 0C F4 01 CA 0C C2 0D F4
0E18 00 CA 0D 90 C0 C4 08 C8
0E20 08 8F FF B8 04 9C FA 3F
0E28 00
```

Once started, the program writes '00' (16) (or 'FF' (16)) into each byte of the section of RAM to be tested. The latter can be selected by entering the start and end addresses of the section of RAM in question. The program then writes the byte '55' (16) into the first ¼ K of the suspect block of RAM (the Elektor 4 K RAM card is structured in blocks of ¼ K).

The program next tests to see whether the byte '55' (16) appears anywhere else in the memory. If that is the case, then it means that there is obviously a fault in the address decoding of the RAM. If the program fails to locate the byte '55' (16) elsewhere in memory, then the program repeats the same test, but this time with the byte '55' (16) in the second ¼ K of RAM to be tested. The cycle is repeated until the entire suspect block of memory has been checked.

When an error *is* detected, this is registered on the displays and the test program is stopped. It can only be continued if the fault in question is rectified.

The next stage in the program consists of loading the byte '00' (16) into every location in the memory and testing to check the contents. If everything is in order, the test is repeated, but this time for the byte 'FF' (16). If all these tests prove negative, then one can safely assume that the malfunction of a program loaded into the area of memory in question is *not* due to a fault in the memory hardware.

Program

The complete condensed listing of the RAM-diagnostic program is given in table 1. The program is loaded from location 0C00 and started from that address. The byte 'FF' (16) for the first test can be loaded into address 0C39.

When the program is started the text 'DG ......' should appear on the displays, whereupon the start address of the section of memory to be tested should be entered. As is apparent from the flow-diagram, this should be the initial address of a ¼ K RAM IC, and the length of the section of memory under test must be a multiple of ¼ K (x 8). The start address should be immediately followed by the end address of the block of suspect memory (e.g. 1000 to 1FFF). The program will start as soon as the end address has been entered. To test an entire 4 K RAM card takes approximately 2 minutes. It is perfectly possible to test larger amounts of memory, thereby crossing page boundaries. The time taken increases exponentially with the size of the memory to be tested. During the test the last entered address will remain on the displays.

The program can terminate in one of four different ways:

'Error no'. When this appears on the displays it indicates that the RAM in question has been given a clean bill of health. This text is always immediately followed by first the start and then the end address of the section of memory which has just been tested.

'Error 1'. For this text to appear the program must have found the byte '55' (16) in the wrong section of memory. This indicates a fault in the address decoding of the memory, caused, for example, by a short in the $\overline{CE}$ line of the RAMs in question. This text is immediately followed by the 'block address' of the faulty ¼ K of RAM. If the precise address of the faulty store is desired, this can be obtained by using the 'MODIFY' routine to load 'FF' into address 0C38 and then restarting the program. The exact address where the error is found will then be displayed immediately after the start and end addresses of the faulty 'block'.

'Error 2'. This text indicates that one or more bits have not been reset (loaded with '0'). The first address where this is the case is registered on the displays, followed by a two-digit number which indicates which bit(s) in the byte are false. For example, '0F' would mean that bits 0, 1, 2 and 3 of the byte in question contained a '1' instead of a '0'. Possible causes for such an error include a faulty RAM IC, bad solder joints, faulty IC sockets, a break in the $\overline{CE}$ or Read/Write lines.

'Error 3'. This is basically the same fault as in the previous case, except that here each bit is tested for a '1'. Once again the first address where a fault is detected is displayed.

In conclusion, it should be remarked that an 'Error' indication need not always denote a fault in the memory hardware. It is perfectly possible that the program will detect an 'error', which if the program is run a second time, will promptly disappear. In such a case the 'fault' is probably due to such factors as, e.g., unsatisfactory supply voltage(s), faulty decoupling capacitors or bus-drivers with too small a fan-out.

reading elektor

TUNs and TUPs

Nowadays, most low-frequency, small-signal silicon transistors from reputable manufacturers meet the following minimum specifications:

$U_{CEO, max}$	20 V
$I_{C, max}$	100 mA
$h_{fe, min}$	100
$P_{tot, max}$	100 mW
$f_{T, min}$	100 MHz

When a transistor of this type is required, it is referred to in Elektor as a **TUN** (Transistor, Universal, NPN) or a **TUP** (Transistor, Universal, PNP).
Some TUNs are the BC 107, BC 108 and BC 109 families, and the 2N3856A, 2N3859, 2N3860, 2N3904, 2N4124 and HEP S0011.
Some TUPs are the BC 177, BC 178 and BC 179 families, and the 2N2412, 2N3251, 2N3906, 2N4126, 2N4291 and HEP S0013.

DUS and DUG

Similarly, for many small-signal applications the only really important difference between all the available diodes is that some are silicon and some are germanium. When a general-purpose small-signal diode is required in an Elektor circuit, it is often designated **DUS** (Diode, Universal, Silicon) or **DUG** (Diode, Universal, Germanium). However, it should be noted that even a DUS or DUG should meet minimum specifications:

	DUS	DUG
$U_{R, max}$	25 V	20 V
$I_{F, max}$	100 mA	35 mA
$I_{R, max}$	1 μA	100 μA
$P_{tot, max}$	250 mW	250 mW
$C_{D, max}$	5 pF	10 pF

Some DUS's are: BA 127, BA 217, BA 218, BA 221, BA 222, BA 317, BA 318, BAX 13, BAY 61, 1N914 and 1N4148.
Some DUGs are: OA85, OA91, OA95 and AA116.

Resistors

Unless otherwise specified, resistors are ¼ Watt 5% tolerance carbon types. Higher power ratings are, of course, permissible (e.g. the recently introduced '¹/₃ Watt' types), provided they fit on the p.c. board — if this is to be used. A 10% tolerance type is usually also permissible, with only a minor effect on the performance.
The resistance values are specified using 'k' for 1000 Ω and 'M' for 1,000,000 Ω; the decimal point is replaced by either 'Ω', 'k' or 'M'.
For instance, $4\Omega7 = 4.7\ \Omega$;
$4k7 = 4700\ \Omega$; $4M7 = 4.7\ M\Omega$.

Capacitors

The DC working voltage of capacitors (other than electrolytics) is normally assumed to be at least 60 V, unless otherwise specified. Generally speaking, of course, a DC working voltage equal to (or greater than) twice the supply voltage is sufficient. In most circuits where electrolytic capacitors are used, a working voltage equal to the supply voltage plus 20% is safe; very often, a lower voltage is sufficient. In recent Elektor circuits, the lowest DC working voltage permissible is often specified, regardless of availability; in practice, any higher voltage type can be used — bearing in mind that a higher voltage rating involves greater physical size, so the available space on the board should be watched. For instance, in a circuit operating off a 9 V battery, an elco might be specified as 1 μ/16 V — even though the normally available types are 1 μ/63 V.
Capacitor values are specified using 'p' for 10^{-12}, 'n' for 10^{-9} and 'μ' for 10^{-6}. As with resistors, the decimal point is replaced by one of these letters. This means that 4700 pF, for instance, is written as 4n7 — not as 0.0047 μF!

Voltages

The international letter symbol 'U' for voltage is normally used instead of the ambiguous 'V'. 'V' is normally reserved for 'volts'. For instance: $U_b = 10\ V$, not $V_b = 10\ V$.
The DC test voltages shown in circuits are measured with a 20 kΩ/V instrument, unless otherwise specified.
No mains (power line) voltages are given in Elektor circuits. It is assumed that our readers know what voltage is standard in their part of the world!

International problems

Although it is the intention that Elektor circuits can be built and used all over the world, some problems are unavoidable.
Readers in countries that use 60 Hz should note that Elektor circuits are designed for 50 Hz operation. This will not normally be a problem; however, in cases where the mains frequency is used for synchronisation some modification may be required. In some cases, the necessary modifications are given; in others, especially where modification involves drastic re-design, the circuit idea may be of use — even though the circuit itself is not.
Circuits intended for use with domestic television receivers may also run into difficulties. As many of our readers will know, 'Television Standards' are anything but 'standard'! However, by avoiding sound and colour wherever possible, a reasonable degree of international compatibility can be achieved — provided the preset adjustments are given a sufficiently wide range.
Finally, especially for our readers in Canada and the USA, a list of equivalents for some of the commonly-used transistor types may prove useful:
AF 239: G0003
BC 140, BC 141, BC 142: S3011
BC 160: S3012
BC 546: no direct equivalent known to us, however in most cases an S0024 can be used
BD 241: TIP29, S5000
BD 242: TIP30, S5006
BF 259, BF 494, BFT 66: no direct equivalent known to us; basically, these are low-voltage, low-current high-frequency devices.
Some further equivalents differ only in that the emitter and collector connections are transposed, so that they must be mounted 'back-to-front' on the EPS p.c. boards:
TUN: RS2010, RS2016
TUP: RS2022, RS2034
BC 107 or equ.: S0015
BC 177 or equ.: S0019
BC 109B,C or equ.: S0024
BC 517: S9100
BC 556: S0026

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TUPTUNDUGDUS

Wherever possible in Elektor circuits, transistors and diodes are simply marked 'TUP' (Transistors, Universal PNP), 'TUN' (Transistor, Universel NPN), 'DUG' (Diode, Universal Germanium) or 'DUS' (Diode, Universal Silicon). This indicates that a large group of similar devices can be used, provided they meet the minimum specifications listed in tables 1a and 1b.

	type	U_{ceo} max	I_c max	h_{fe} min.	P_{tot} max	f_T min.
TUN	NPN	20 V	100 mA	100	100 mW	100 MHz
TUP	PNP	20 V	100 mA	100	100 mW	100 MHz

Table 1a. Minimum specifications for TUP and TUN.

Table 1b. Minimum specifications for DUS and DUG.

	type	U_R max	I_F max	I_R max	P_{tot} max	C_D max
DUS	Si	25 V	100 mA	1 µA	250 mW	5 pF
DUG	Ge	20 V	35 mA	100 µA	250 mW	10 pF

Table 2. Various transistor types that meet the TUN specifications.

TUN		
BC 107	BC 208	BC 384
BC 108	BC 209	BC 407
BC 109	BC 237	BC 408
BC 147	BC 238	BC 409
BC 148	BC 239	BC 413
BC 149	BC 317	BC 414
BC 171	BC 318	BC 547
BC 172	BC 319	BC 548
BC 173	BC 347	BC 549
BC 182	BC 348	BC 582
BC 183	BC 349	BC 583
BC 184	BC 382	BC 584
BC 207	BC 383	

Table 3. Various transistor types that meet the TUP specifications.

TUP		
BC 157	BC 253	BC 352
BC 158	BC 261	BC 415
BC 177	BC 262	BC 416
BC 178	BC 263	BC 417
BC 204	BC 307	BC 418
BC 205	BC 308	BC 419
BC 206	BC 309	BC 512
BC 212	BC 320	BC 513
BC 213	BC 321	BC 514
BC 214	BC 322	BC 557
BC 251	BC 350	BC 558
BC 252	BC 351	BC 559

Table 4. Various diodes that meet the DUS or DUG specifications.

DUS		DUG
BA 127	BA 318	OA 85
BA 217	BAX 13	OA 91
BA 218	BAY 61	OA 95
BA 221	1N914	AA 116
BA 222	1N4148	
BA 317		

Table 5. Minimum specifications for the BC107, -108, -109 and BC177, -178, -179 families (according to the Pro-Electron standard). Note that the BC179 does not necessarily meet the TUP specification ($I_{c,max}$ = 50 mA).

		NPN	PNP
		BC 107	BC 177
		BC 108	BC 178
		BC 109	BC 179
U_{ceo}	max	45 V	45 V
		20 V	25 V
		20 V	20 V
U_{ebo}	max	6 V	5 V
		5 V	5 V
		5 V	5 V
I_c	max	100 mA	100 mA
		100 mA	100 mA
		100 mA	50 mA
$P_{tot.}$	max	300 mW	300 mW
		300 mW	300 mW
		300 mW	300 mW
f_T	min.	150 MHz	130 MHz
		150 MHz	130 MHz
		150 MHz	130 MHz
F	max	10 dB	10 dB
		10 dB	10 dB
		4 dB	4 dB

The letters after the type number denote the current gain:

A: α' (β, h_{fe}) = 125-260
B: α' = 240-500
C: α' = 450-900.

Table 6. Various equivalents for the BC107, -108, ... families. The data are those given by the Pro-Electron standard; individual manufacturers will sometimes give better specifications for their own products.

NPN	PNP	Case	Remarks
BC 107 BC 108 BC 109	BC 177 BC 178 BC 179		
BC 147 BC 148 BC 149	BC 157 BC 158 BC 159		P_{max} = 250 mW
BC 207 BC 208 BC 209	BC 204 BC 205 BC 206		
BC 237 BC 238 BC 239	BC 307 BC 308 BC 309		
BC 317 BC 318 BC 319	BC 320 BC 321 BC 322		I_{cmax} = 150 mA
BC 347 BC 348 BC 349	BC 350 BC 351 BC 352		
BC 407 BC 408 BC 409	BC 417 BC 418 BC 419		P_{max} = 250 mW
BC 547 BC 548 BC 549	BC 557 BC 558 BC 559		P_{max} = 500 mW
BC 167 BC 168 BC 169	BC 257 BC 258 BC 259		169/259 I_{cmax} = 50 mA
BC 171 BC 172 BC 173	BC 251 BC 252 BC 253		251...253 low noise
BC 182 BC 183 BC 184	BC 212 BC 213 BC 214		I_{cmax} = 200 mA
BC 582 BC 583 BC 584	BC 512 BC 513 BC 514		I_{cmax} = 200 mA
BC 414 BC 414 BC 414	BC 416 BC 416 BC 416		low noise
BC 413 BC 413	BC 415 BC 415		low noise
BC 382 BC 383 BC 384			
BC 437 BC 438 BC 439			P_{max} = 220 mW
BC 467 BC 468 BC 469			P_{max} = 220 mW
	BC 261 BC 262 BC 263		low noise

variable fuzz-box

A simple circuit for musical sound effects

Particularly in modern pop music, electronically-produced sound effects are an extremely common device. One only has to think, for example, of the widespread use of the 'wa-wa' pedal and the fuzz-box. With this in mind, Elektor have designed a super simple circuit which, by employing signal clipping techniques, can produce a large variety of manually controlled sound effects.

Using only a handful of components it is possible to construct a highly effective variant of the well-known fuzz box. This type of circuit commonly employs a pair of anti-parallel connected diodes, which are inserted in the amplifier (be it IC — or transistorised) so as to clamp the output signal above certain values of the input signal. This process is illustrated in figure 1, where for the sake of clarity the amplifier is assumed to have unity gain over the linear portion of its transfer characteristic. As can be seen,

above an input voltage U_1, the output voltage shows no further increase; similarly, the output voltage will not fall below the input value $-U_2$. If U_1 is equal to U_2 (which is typically the case in fuzz-box circuits) and the input signal is sufficiently large, then the input and output signals will differ from one another as shown in figure 1b.

When, as is the case in figure 1b, the output signal clips *symmetrically* (i.e. $U_1 = U_2$), it contains only *even* harmonics, and it is this which gives the

Figure 1. The response of an amplifier which is driven into 'hard' clipping (a) and the corresponding input and output waveforms.

Figure 2. There are five parameters of the clipping response of an amplifier which can be independently varied.

3

Figure 3. The circuit diagram of a variable fuzz-box for a symmetrical power supply.

Figure 4. The circuit of a variable fuzz-box for an asymmetrical power supply.

Figure 5. Example of a suitable power supply for the circuit of figure 3 (a) and of figure 4 (b).

4

resultant musical signal its characteristically rough and 'fuzzy' edge. However the tonal character of the music signal can be considerably enriched by arranging that the output signal clips *asymmetrically* (i.e. $U_1 \neq U_2$). In this way it is possible to influence the sound of the fuzz-box to produce more varied effects. The circuit described here is designed to offer the best of both worlds by allowing the clipping levels U_1 and U_2 to be altered independently of one another, thus permitting the type of effect to be varied as desired.

Varying the clipping levels is not, however, the only way of influencing the sound of the (clipped) output signal. A further factor determining the type of effect produced is whether the amplifier starts clipping almost immediately it reaches a particular level (hard clipping), or whether the transition between non-limiting and limiting is more gradual (soft clipping). In the circuit described here it is possible to continuously vary the response of the amplifier between these two extremes. The various control facilities offered by the 'variable fuzz-box' are illustrated in figure 2.

Circuit diagram

The complete circuit diagram of the fuzz-box is shown in figures 3 and 4; figure 3 gives the circuit for a symmetrical power supply (plus, minus and earth), whilst figure 4 is designed for an

5a

BD 135
BD 137

11,5 ... 14,5 V

F1
250 mA

1000µ
40V

1k5

220µ
16V

12...15V
400mW

100n

12...15V
400mW

220µ
16V

100n

1000µ
40V

1k5

2x 12 ... 18 V
50 mA

4 x 1N4001

B

BD 136
BD 138

11,5 ... 14,5 V

9984 - 5a

5b

BD 137
BD 139

40V (60 V)

21,5V (32,5 V)

10mA

1k8
(2k7)
½ W

100n

22V (33V)
400mW

220µ
40V

9984 - 5b

asymmetrical supply (plus and minus/ earth). The current consumption of the circuit is 10 to 15 mA, whilst in both figures 3 and 4 the input impedance is 100 k.

The operation of the circuit is fairly straightforward. The input signal is first amplified by IC1, the gain of which is $1 + \frac{R2}{R1}$. The gain can be altered by varying R1; with the value given the gain is 11. The output signal of IC1 is fed via R4 (and C6 in figure 4) to the volume control P5, the wiper voltage of which forms the output signal of the circuit.

The clipping is controlled as follows: as soon as the voltage at the right hand end of R4 exceeds the wiper voltage of P3 (or falls below the wiper voltage of P4) the output signal is attenuated. The degree of attenuation is determined by the ratio of R4 to P1 (P2). With P1 (P2) set to its minimum resistance value the output signal is completely attenuated, i.e. is 'hard' clipped. With P1 set to its other end stop (minimum resistance) 'soft' clipping is obtained. Thus by adjusting the four potentiometers P1 ... P4 which control both the levels at which the amplifier starts to clip and the degree of clipping, it is possible to vary the tonal character of the resultant sound as desired.

As far as a power supply is concerned, various possibilities exist. Figures 5a and 5b show a suitable arrangement for the circuits of figures 3 and 4 respectively.

Other applications

Apart from its use as a variable special effects generator the circuit also has other possible applications. For example, it can be used to limit the input signal of a power amplifier to the value which just produces maximum output. In this way one obviates the danger of current limiting or clipping occurring in the power amp and its undesirable consequences for the listener. At the same time the circuit thus represents a type of overload protection for the power amp.

Another interesting idea would be to use the circuit in conjunction with P.A. systems. There are various theories which claim that it is possible to increase the volume of the signal by clipping it in a certain way. The increase in the harmonic content of the signal is said not to impair the intelligibility of the address. It has also been suggested that different clipping characteristics may explain some of the oft-debated differences between 'transistor-sound' and 'valve-sound'. The circuit described here seems ideally suited to test the truth of these ideas. However readers wishing to experiment in this field would do well to spare a thought for their neighbours!
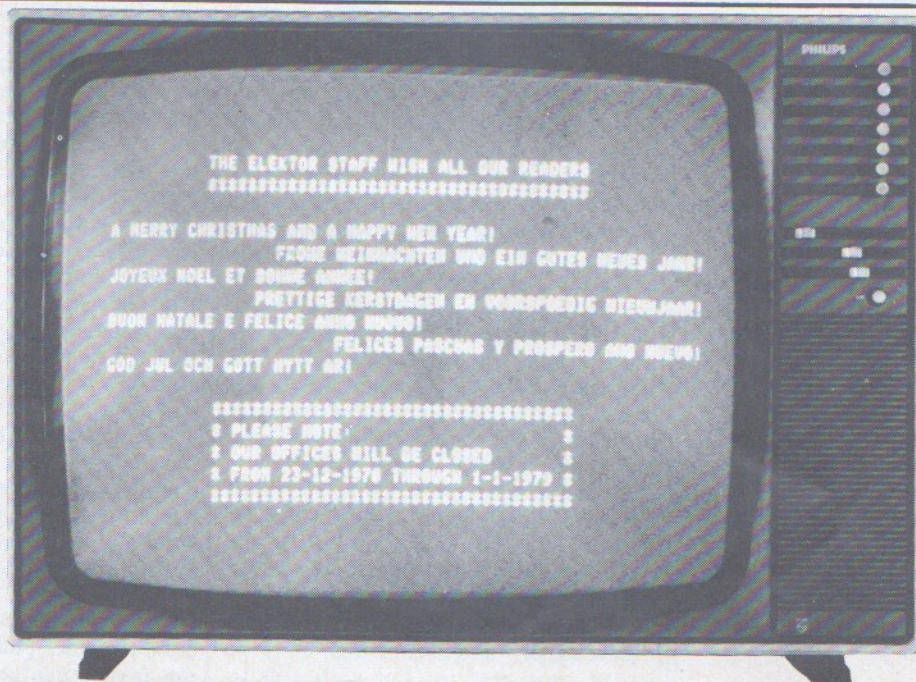
TV scope - extended version

The TV scope, introduced in the recent October issue, can now be completed. Two months ago the basic version was described in detail and the necessary circuits were given for converting a standard domestic television receiver into an oscilloscope — be it with limited capabilities. Last month the possibilities for extending the TV scope were discussed; in particular, the principles and advantages of a variable timebase and trigger facilities were explained. This third and final article in the series gives the necessary extension circuits.

Although the basic version is an extremely useful tool when testing low-frequency circuits, its 'big brother' offers vastly superior display possibilities. Any signal within the frequency range of the scope (DC to 100 kHz) can be displayed as a clear and stable trace on the screen of any domestic TV receiver.

Several block diagrams can be given for the same circuit, as shown in figure 1: the third (and last!) block diagram of the 'TV scope - extended version'. The sections required for extending the TV scope are shaded in; the unshaded portions are the basic version described in the October issue. Furthermore, the block diagram clearly shows the separate portions of the circuit which will be mounted on separate printed circuit boards. In the following, these extension circuits will be discussed: the trigger circuit, the input timebase and the 'bucket-brigade memory' timebase expander with its associated drive circuits. The portions of the diagram shown in dotted lines are only required for a two-channel version of the 'scope'. The timebase expanders (bucket-brigade memories) of the A- and B-channels (assuming that the latter is also included) 'slow down' the analogue input signals u_{ya} and u_{yb} as required. For correct operation, they require three control signals: two clock signals, $\emptyset 1$ and $\emptyset 2$, and a 'mode' signal (u_m) that determines which of the memories is in the 'read' cycle and which is in the 'display' mode. The basic principles involved were described last month.

The three control signals are derived from a fairly complicated logical circuit, which is designated quite simply in the block diagram: the 'control circuit'. This circuit derives the necessary output signals from several input signals. Four inputs are provided by the main board of the basic version of the TV scope. Two further inputs are provided by the trigger circuit, which forms part of the extension circuitry. One of the latter signals, u_{sample}, corresponds to twice the clock frequency required during the 'read' cycle of the memories. The frequency of u_{sample} therefore determines the timebase expansion obtained; in other words, it determines the time axis of the final TV scope display.

The second signal provided by the input timebase board is u_x. This signal is derived from the $\overline{u_{reset}}$ output of the basic version: a pulse train at the 50 Hz frame frequency. The u_x signal is also a pulse train — in other words, it is a digital signal — and it goes 'high' (+15 V) in the same rhythm, but with a delay

with respect to $\overline{u_{reset}}$ that is determined by the setting of a potentiometer, 'x-pos'. The u_x signal determines the start of the display cycle with respect to the start of each frame of the TV picture. In effect, it therefore determines the position of the display along the time axis or 'X-axis', which is why the corresponding control is labelled 'x-pos(ition)'.

The last input signal to the control circuit is u_{tr}. This signal is derived from one of the y-input signals or from a third input signal ('external trigger'). When the selected input signal to the trigger circuit exceeds a certain level, as set by the 'trigger level' control, the u_{tr} output goes 'high' — see figure 2. Since the positive-going edge is detected by the control circuit, triggering occurs on the leading edge of the selected input signal. Basically, this circuit is equivalent to similar circuits in any 'normal' oscilloscope.

Figure 3 shows the relationships between the various in- and output signals of the control circuit. At each reset pulse, the mode signal (u_m) changes state, selecting first one memory and then the other for each channel. Simultaneously the u_x signal goes low for a fixed period; at the end of this period the display cycle starts. Two things happen at this point. The signal u_{gate} goes high; this signal is fed to one input of N15 in the circuit which generates the white-level pulses for the trace on the TV screen (see the description of the basic version, figure 4c), so that a trace only appears on the TV screen when one of the memories is actually in its display cycle. Between display cycles, the trace is blanked.

The second signal initiated by the positive-going edge of u_x is the display clock pulse train. As explained last month, this signal must be fed to the two memories alternately — each in turn being used as display memory — and for this reason these (relatively low frequency) clock pulses appear in alternate bursts on the $\emptyset 1$ and $\emptyset 2$ lines.

When one memory (for each channel) is in the display mode, the other is 'storing' the input signal. This 'read' cycle is initiated by the first trigger pulse, u_{tr}, after the reset pulse. A control signal to the input timebase, u_{et} ('enable time-

base'), goes positive. The input timebase clock generator starts on the positive-going edge of u_{et} and produces a rapid burst of clock pulses, alternately on the Ø1 and Ø2 lines. As in a conventional scope, this facility is included in order to obtain a stable display — independent of input frequency.

Having surveyed the basic principles involved, it is now time to discuss the necessary circuits in greater detail.

Trigger circuit

The trigger circuit is shown in figure 4. A three-way switch selects the desired trigger input: either of the two input signals to the TV scope, or an independent trigger input. If either of the two

Y-inputs to the scope is selected, a signal at 'standard level' is required. This signal, u_t, is derived from the input amplifier(s) of the TV scope — see 'TV scope - basic version', figure 2 on page 10-31.

Depending on the type of signal selected, either AC or DC coupling may give better results. Switch S2, bridging C1, is used to select the desired option. IC1 boosts the input signal to a level suitable for the comparator stage, IC2. This comparator refers the input signal level to a DC voltage set by the 'trigger level' potentiometer. If the (amplified) input signal exceeds the preset level, the output of IC2 goes high (see figure 2) — note that IC1 inverts the input signal!

As long as trigger pulses are appearing at regular intervals, LED D2 lights ('trig'd'). The output signal, u_{tr}, is fed to the control circuit on the memory board.

The printed circuit board for the trigger circuit contains one further control: potentiometer P2, 'trace distance'. This control has nothing to do with the trigger circuit, but it happened to fit neatly at this point on the front panel. The function of this control will be described further on.

The trigger circuit and the trace distance control are both mounted on the board shown in figure 5. The 'AC/DC' switch S2 and capacitor C1 are mounted off-board. Basically, of course, any LED can be used for D2. However, the neatest
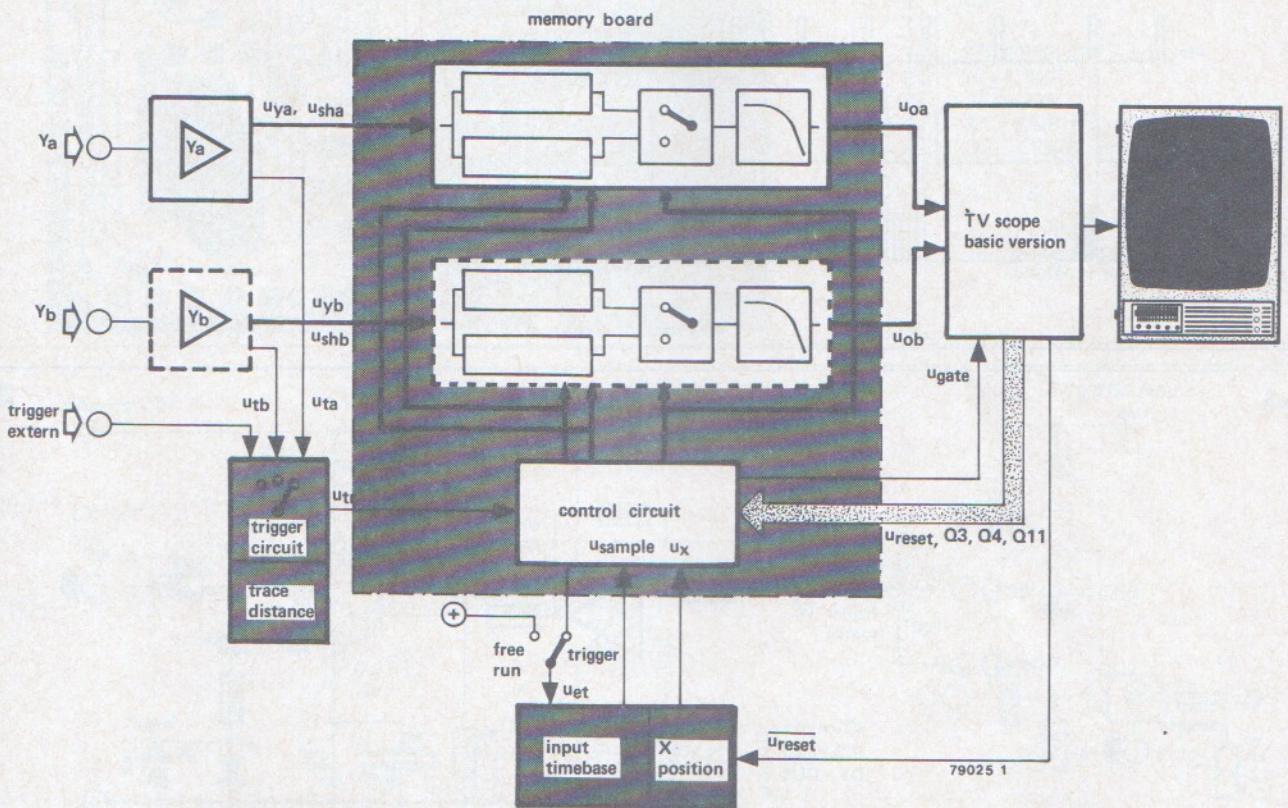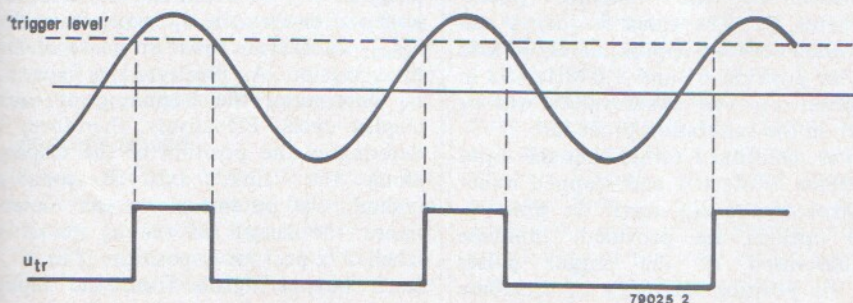

Figure 1. Block diagram of the TV scope — extended version. The shaded sections are the extension circuits.

Figure 2. Trigger pulses (u_{tr}) occur when the 'trigger source' signal exceeds the 'trigger level'.

3

Signal waveforms labelled (top to bottom):
u_m, u_{reset}, u_x, u_{gate}, $\phi 1$, u_{tr}, u_{et}, $\phi 2$
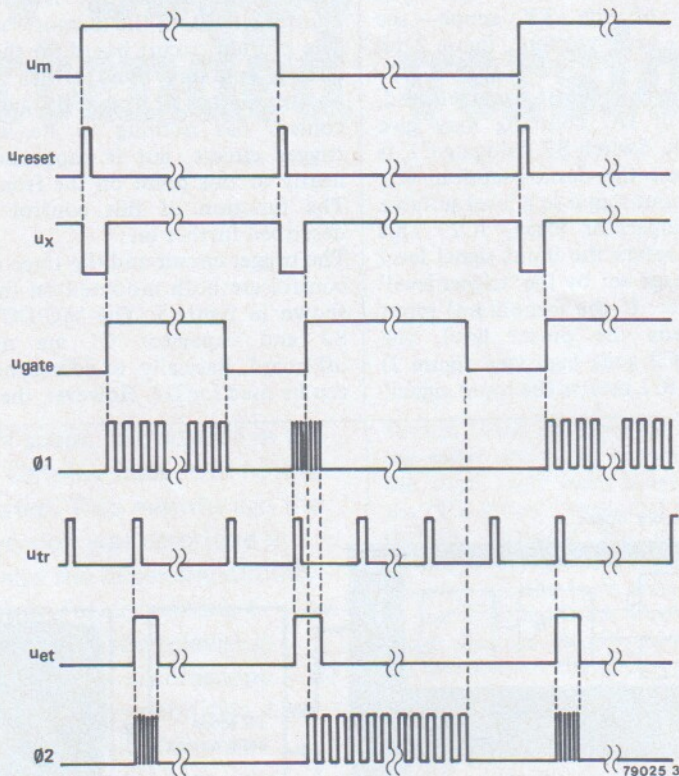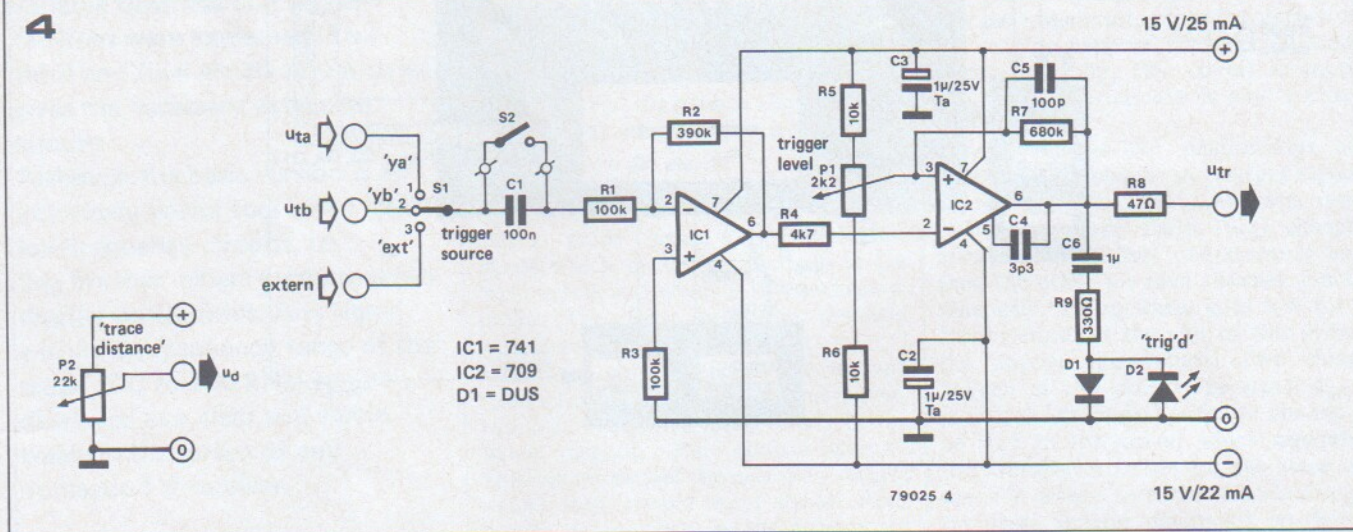
79025 3

Figure 3. The relationships between the various in- and output signals of the control circuit.

Figure 4. The trigger circuit, which derives the u_{tr} pulses from the analogue 'trigger source' input. The corresponding printed circuit board also contains the 'trace distance' control, P2, since it happens to fit neatly at this point on the front panel.

Figure 5. The 'trigger board' (EPS 9969-2). All components shown in figure 4, with the exception of S2 and C1, are mounted on this board; the miniature LED, D2, is mounted on the copper side of the board.

Figure 6. The input timebase is a CMOS square-wave generator; the frequency is determined by a selector switch and a fine frequency control. The remaining gate in the same IC, N4, is used to derive the u_x pulses from the u_{reset} signal.

4

IC1 = 741
IC2 = 709
D1 = DUS

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Figure 4 circuit labels: u_{ta}, u_{tb}, extern, 'ya', 'yb', 'ext', S1, S2, C1 100n, trigger source, R1 100k, IC1, R2 390k, R3 100k, R4 4k7, trigger level, P1 2k2, R5 10k, R6 10k, C2 1µ/25V Ta, C3 1µ/25V Ta, IC2, C4 3p3, R7 680k, C5 100p, R8 47Ω, C6 1µ, R9 330Ω, D1, D2, 'trig'd', u_{tr}, 'trace distance', P2 22k, u_d, 15 V/25 mA, 15 V/22 mA.

result is obtained if a 3 mm Ø green LED is used; if it is mounted on the copper side of the board it will fit in the corresponding hole in the front panel.

Input timebase
The input timebase is the generator for the input clock signal, u_{sample}. The frequency of this signal determines the sampling rate of the input signal(s) to the TV scope. As described earlier, u_{sample} is a square-wave signal at twice the frequency required for sampling and storing the input signal(s).
The circuit of the input timebase is given in figure 6. The upper portion of this diagram is the clock generator proper. This is a CMOS multivibrator,

the frequency of which can be selected by means of the six-way switch S2; fine control of the output frequency is provided by the 'time/div' potentiometer, P1. The clock frequency can be varied between approximately 32 kHz (S2 in position 6) and 1.6 MHz (S2 in position 1). The exact values will be given in the calibration procedure.
It was mentioned earlier that the input timebase is started and stopped under control of the u_{et} signal. In practice, two options are provided: timebase synchronised to the trigger pulses (S1 in position 'trigger') or timebase running continuously (S1 in position 'free run').
As in the case of the trigger circuit, the

'input timebase' printed circuit board actually contains a further circuit that has nothing to do with the timebase proper. N4, with its associated components, derives the u_x signal from the $\overline{u_{reset}}$ signal from the main board of the basic version. As illustrated in figure 3, u_x determines the beginning of each display cycle. Effectively, therefore, it determines the position of the display along the 'time' axis. Reasonably enough, the potentiometer that determines the length of the u_x pulse is labelled 'x-pos' for 'x-position' (P8).
Both output signals from the input timebase p.c. board are fed to the 'memory board'.
A printed circuit board and component

5
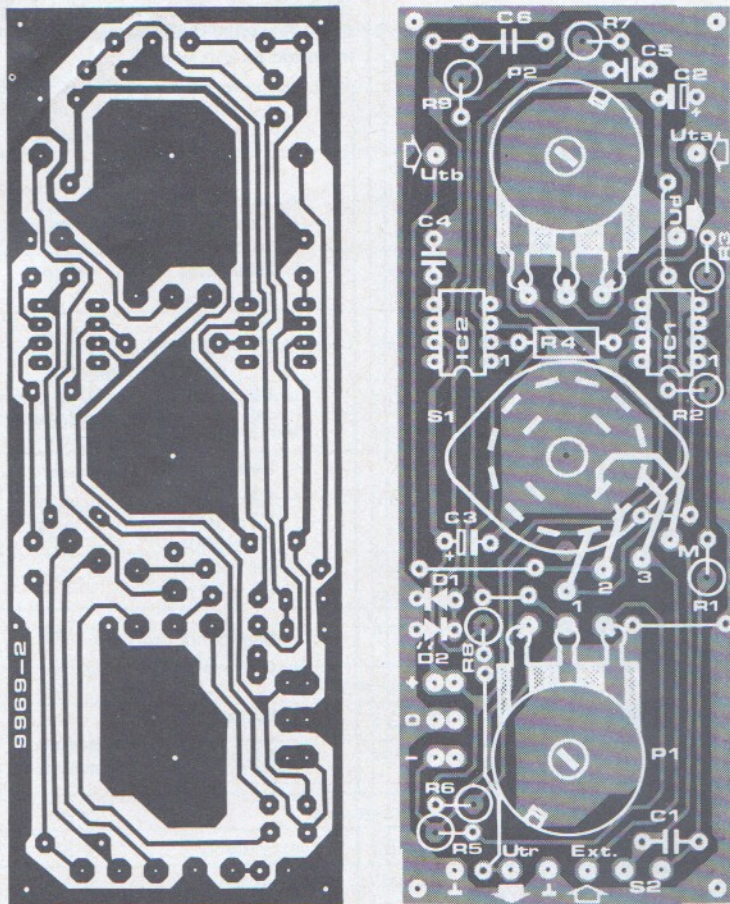
Parts list trigger circuit

Resistors:

R1,R3 = 100 k
R2 = 390 k
R4 = 4k7
R5,R6 = 10 k
R7 = 680 k
R8 = 47 Ω
R9 = 330 Ω
P1 = 2k2 (2k5) lin
    potentiometer
P2 = 22 k (25 k) lin
    potentiometer

Capacitors:

C1 = 100 n
C2,C3 = 1 μ/25 V tantalum
C4 = 3p3
C5 = 100 p
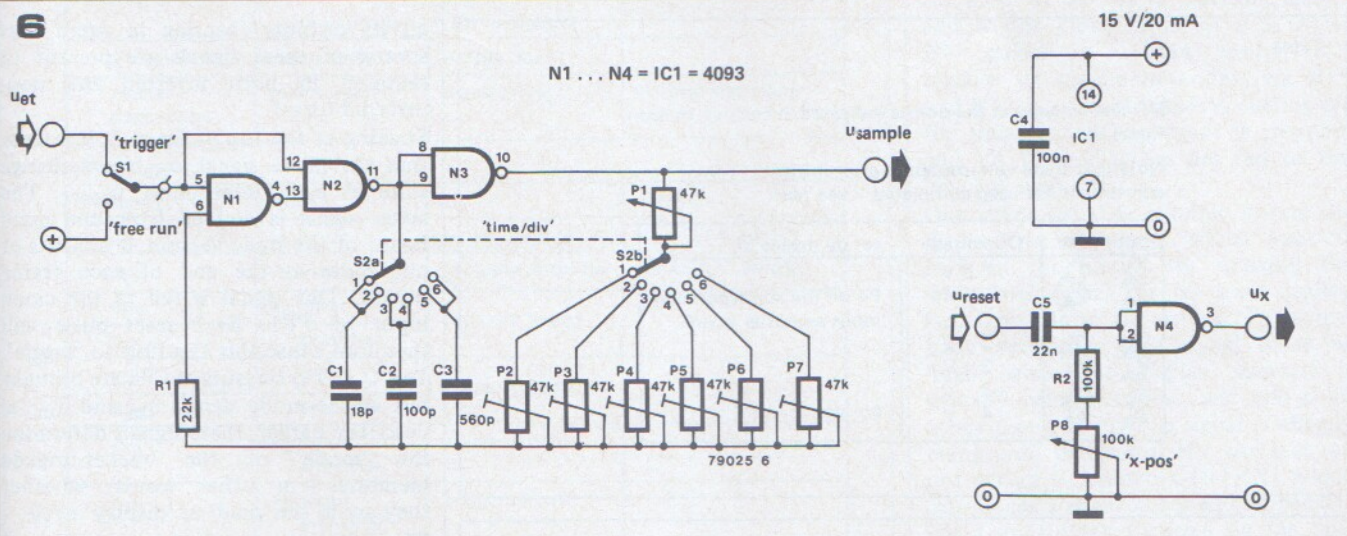C6 = 1 μ

Semiconductors:

IC1 = 741
IC2 = 709
D1 = DUS
D2 = LED (3 mm, green)

Sundries:

S1 = 1-pole, 3-way switch
S2 = single-pole,
    single-throw switch

6

layout for the input timebase and x-position circuit are given in figure 7. Switch S1 is mounted off-board. The adjustment procedure for the six preset potentiometers will be described later. Meanwhile, as with almost any other circuit, it is advisable to set all presets to the mid-position for the time being.

Memory board

It was already fairly clear from the block diagram shown in figure 1: most of the extension circuits for the TV scope are contained on one board. The 'memory board'. Two circuits are mounted on this p.c. board: the control circuit (figure 8a) for the memories, and the bucket-brigade memories with their associated circuitry (figure 8b).

The control circuit is a fairly straightforward logic circuit using CMOS ICs. Like most logic circuits (and in contrast to most analogue circuits) it receives a multitude of input signals and produces a similar plethora of outputs. From the main board (basic version) it receives three timing signals, Q_3, Q_4 and Q_{11}, and u_{reset} (both in normal and in inverted format); the input timebase provides both u_x and u_{sample}, determining the start of the display and the sample-rate of the store cycles, respectively; finally, the trigger circuit provides a signal, u_{tr}, which determines the start of the store cycle.

From these input signals, the control circuit derives several output signals. The clock signals for the bucket-brigade memories, $\emptyset 1$ and $\emptyset 2$, both non-inverted and inverted ($\overline{\emptyset 1}$ and $\overline{\emptyset 2}$) and the mode signal, u_m, plus its inverted form ($\overline{u_m}$). In addition to these six control signals for the memories, two further outputs must be obtained: u_{gate}, which controls the white-level pulse gate on the main board, as described earlier; and u_{et}, which enables the clock generator on the input timebase board.

Operation of this part of the extension circuit can be considerably simplified by referring to the pulse diagram shown in figure 3. This diagram shows the required relationships between half the input signals to the control circuit and
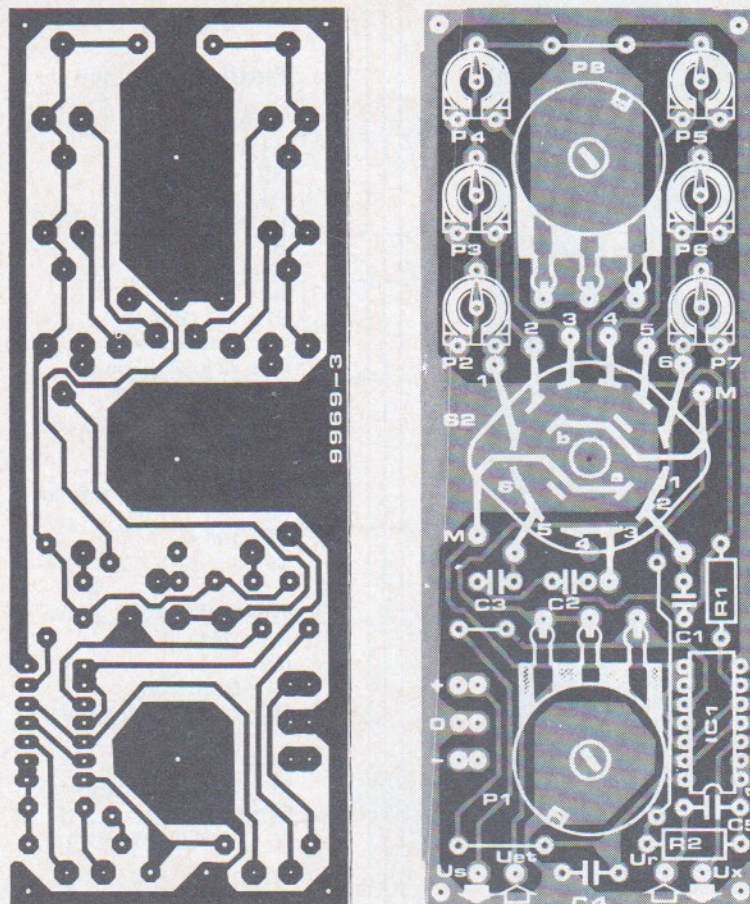
Figure 7. The 'input timebase board' (EPS 9969-3). The components shown in figure 6, with the exception of S1, are mounted on this board.

Parts list input timebase

Resistors:
R1 = 22 k
R2 = 100 k
P1 = 47 k (50 k) lin
     potentiometer
P2 . . . P7 = 47 k (50 k) preset
P8 = 100 k lin potentiometer

Capacitors:
C1 = 18 p
C2 = 100 p
C3 = 560 p
C4 = 100 n
C5 = 22 n

Semiconductors:
IC1 = CD 4093

Sundries:
S1 = 1-pole, 2-way switch
S2 = 2-pole, 6-way switch

Table 1

DC test voltages at the points indicated in figure 8b

Note that these voltages apply after initial adjustment has been completed — see text.

test point	DC voltage	set by means of
①, ①'	5.5 V	
②, ②'	5.5 V	P4 on the corresponding input amplifier board
③, ③'	3 V	
④, ④'	3 V	
⑤, ⑤'	−3 V	
⑥, ⑥'	4 V	P2, P2'

Table 2

Modifications to the basic version

circuit	board no.	figure no. s	component	original value	new value
input amplifier	9968-1	2 & 3, pages 10-31 & 10-32	R10	22 k	5k6
			R23	220 k	2M2
main board	9968-2	4c & 6, pages 10-33 & 10-34	R7	470 Ω	220 Ω
			remove the wire link marked between u_g and +		

all its outputs (bearing in mind that several of these signals are present or required in both inverted and non-inverted form).

Starting at the top of figure 3, it is clear that the mode signal, u_m, must change state at each reset pulse, u_{reset}. The latter signal is derived from the main board of the basic version; it consists of one pulse at the end of each frame period. This signal is fed to the clock input of FF1. Each reset pulse will therefore cause this flip-flop to 'toggle'. The Q and $\bar{Q}$ outputs of FF1 are brought out as the mode signals u_m and $\overline{u_m}$; as described earlier, these signals determine the 'mode' of the bucket-brigade memories — in other words, whether they are in the 'read' or 'display' cycle.

The following signals shown in figure 3 are related to the 'display' cycle. When u_x goes positive (the delay with respect to the u_{reset} pulse being determined by the setting of the 'x-pos' potentiometer on the input timebase board) the display cycle must start. To understand the operation of the control circuit during this cycle, we must first side-track briefly and consider the control signals Q_3 and Q_4 and the function of IC5.

IC5 (and IC3, of which more later) is a 14-bit binary counter. Only 8 bits are actually used in this circuit, the output being taken from the Q_9 pin, so that the maximum count corresponds to 256 clock pulses. These pulses are derived, via N9 and S1, from either the Q_3 or the Q_4 output of the main board.

Table 3

Interconnections between boards for the extended version

insulated wire

signal/voltage	from	to	remarks
u_{sync}			
u_{cal}	main board	video mixer	
u_w			
u_{tb}/u_f	main board	main board	wire link on main board
u_x	input timebase	memory board	
u_{sample}			
$\overline{u_{reset}}$	main board	input timebase	
u_{reset}	main board	memory board	
u_{gate}			
u_{reset}	main board	memory board	
Q_{11}			
u_{et}	input timebase	'free run' switch	
u_{et}	'free run' switch	memory board	
0 V	all boards		
0 V	video mixer	LED D3	
0 V	mains trafo	supply	case connected to supply common
+15 V	all boards		note R_a, R_b, C_a, C_b
+15 V	video mixer	LED D3	via R7
+15 V	input timebase	'free run' switch	
−15 V	all boards, with exception of TV modulator		
2 x 18 V	mains trafo	supply board	
mains connection	mains input	transformer	via fuse and mains switch, twisted

screened cable

signal	from	to	remarks
u_{ia}	y_a input	y_a input amplifier	via AC/DC switch
u_{ib}	y_b input	y_b input amplifier	via AC/DC switch
u_{ya}	y_a input	memory board	
u_{sha}	amplifier		
u_{yb}	y_b input	memory board	
u_{shb}	amplifier		
u_{ta}	y_a input amplifier	trigger circuit	
u_{tb}	y_b input amplifier	trigger circuit	
u_d	trigger circuit	memory board	
u_{tr}	trigger circuit	memory board	
'ext. trigger'	ext. trigg. input	trigger circuit	
S2 (AC/DC trigger)	trigger circuit	switch	two connections, see figure 10
u_{oa}/u_{ya}	memory board	main board	
u_{ob}/u_{yb}			
u_{video}	video mixer	tv modulator	
u_{video}	tv modulator	video output	
VHF/UHF	tv modulator	VHF/UHF output	
Q_3	main board	'x-magn.' switch	
Q_4			
Q_3/Q_4	'x-magn.' switch	memory board	

With S1 in position 1 ('x magnitude' is 'x 1'), the Q_3 output is selected. The frequency of the pulses at this output corresponds to the frequency of the sawtooth voltage u_{ref} on the main board. The result is that the display cycle of the selected memory, the count in IC5, and the final display on the TV screen all run in step: each sample value in the memory is displayed at one point on the screen. When S1 is placed in position 2, however, the Q_4 output is selected — the frequency of which is half that of the u_{ref} sawtooth. In this case, the memory is read out at half the speed of the display sweep, so that each value stored in memory is actually displayed as a dot on two consecutive lines in the picture. This results in a final display that is 'stretched' along the time axis.

During each display cycle, the selected pulses (either from Q_3 of Q_4) are passed through N7/N14 or N8/N15 — depending on the logic levels of u_m and $\overline{u}_m$ — to either the Ø1 or Ø2 output. The inverted signals, which are also required for clocking the bucket-brigade memories, are obtained from N19 and N20.

Going back now to the u_x input, the operation of the control circuit during the display cycle can be clarified. Initially, u_x was low, holding IC5 in the reset condition and blocking the Q_3 or Q_4 pulses at N9. When u_x goes high, N9 is freed and passes the pulses — both to the Ø1 or Ø2 output, as described above, and to the clock input of IC5. After 256 pulses have been passed, IC5 reaches its maximum count; via N18 and N10 it again blocks N9, cutting off the pulse train. The count is stopped, and no further pulses are fed to the 'display' memory.

One other signal is required during the display period: u_{gate} which controls N15 in the circuit for deriving the white-level pulses for the trace display (see figure 4c in the article 'TV scope, basic version'). This signal must be 'high' when the display memory is actually being read out, and low at all other times. A suitable signal is already being used to control N9, and it is fed out via an RC-network (R1/C1) which introduces a slight delay to compensate for a similar delay caused by the low-pass filters at the output of the memory. When one memory is in the display cycle, the other is sampling the input signal. This 'read cycle' is similar in many ways to the display cycle. A pulse at the $\overline{u_{reset}}$ input resets the RS flip-flop N5/N6 (the Q output goes 'low'); the u_{et} output is then also low. The Q_{11} signal from the main board is also low after the reset pulse, so the output of N1 is high, enabling N3. A positive-going edge at the u_{tr} input is differentiated by C2 and R2, causing the output of N3 to go 'high' briefly. This, in turn, sets the flip-flop (N5/N6); in other words, its Q output goes 'high' on the first positive edge occurring at the u_{tr} input after the reset pulse. This causes

8a

Q3 '1x' 'x-magn'
Q4 S1 Ø3/Ø4
'2x'
u_x — N16 — R — IC5 4020 'Q9' — N18 — N10 — N9 — R1 33k — C1 10n — u_gate / u_m / $\overline{u_m}$

u_sample — CK FF2 Q / D $\overline{Q}$

u_reset — CK FF1 Q / D R S $\overline{Q}$

IC3 4020 'Q9' — N17

$\overline{u_{reset}}$

Q11 — N1 — N3 — N4 — N5 $\overline{Q}$ / N6 Q — N11 — u_et
u_tr — C2 47p — N2 — R2 22k

N12 / N7 — N14 — Ø1
N19 — $\overline{Ø1}$
N13 / N8 — N15 — Ø2
N20 — $\overline{Ø2}$

N1 . . . N4 = IC1 = 4011
N5 . . . N8 = IC2 = 4011
FF1 . . . FF2 = IC4 = 4013
N9 . . . N11 = IC6 = 4081
N12 . . . N15 = IC7 = 4011
N16 . . . N20 = IC8 = 4049

15V + IC8 4049 / C7 100n IC7 4011 / IC6 4081 / C6 22n IC5 4020 / C5 100n IC4 4013 / IC3 4020 / C4 22n IC2 4011 / C3 22n IC1 4011

79025 8a

the output of N11 (the u_et signal) to go high, starting the input timebase oscillator (figure 6). The output from the latter, u_sample, is passed through a flip/flop (FF2) to obtain the input clock pulses. Depending on the mode signal, this pulse train is either fed to the Ø1 or Ø2 output to clock the memory which is in the 'read' mode. Simultaneously, the pulses are counted by IC3; when this reaches its maximum count of 256, its output goes high. This causes the u_et output to go low, stopping the timebase oscillator and, with that, the 'read' cycle.

In the absence of trigger pulses, no 'read' cycle would normally occur. However, approximately halfway through the total period Q_{11} goes high. This has the same effect as a trigger pulse, enabling the input timebase to initiate a read cycle. Effectively, the scope switches over to a 'free run' mode in the absence of trigger pulses.

So much for the control circuit. Not surprisingly, the second part of the 'memory board' contains the memories (figure 8b). The portions enclosed within dotted lines are identical: the upper section is for the A-channel and the lower for the y_b input. The latter section can, of course, be omitted if only a single-channel scope is required.

The heart of the circuit is IC9. This contains two independent bucket-brigade delay lines, each with 512 'buckets'. Clock signals Ø1 and $\overline{Ø1}$ are fed to one of these delay lines; Ø2 and $\overline{Ø2}$ are used to clock the other. In
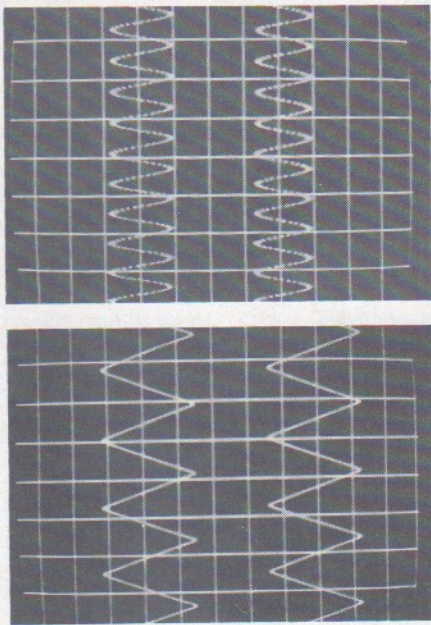
Figure 8. Two closely related circuits are mounted on the 'memory board': the control circuit (figure 8a) and the memory circuit proper (figure 8b). For a two-channel version of the scope, two memory circuits are required; in figure 8b both of these circuits are shown in full.

Principle, both delay lines could receive the same input signal; in practice, however, u_sh (for 'shifted) is not quite identical to u_y. As is apparent from the circuit of the input amplifier (figure 2 in the article dealing with the basic version – see page 10-31 in the October issue), both the signal amplitude and the superimposed DC level of the u_sh signal can be varied slightly with respect to u_y. This is done to compensate for any minor differences between the two delay lines, as will be described for the calibration procedure.

The two delay lines in IC9 each have two outputs. Mixing these two outputs in a resistor network helps to suppress the clock components, providing a total analogue output (at point 3 or 4, respectively) that is relatively 'clean'. Under control from the 'mode' signals (u_m and $\overline{u_m}$), the analogue switches contained in IC10 select one of the two memory outputs for display purposes. The 'unwanted' output is connected unceremoniously to ground.

The selected output is fed through a sixth-order low-pass filter, A1 . . . A3, with a turnover frequency of 2.5 kHz. This frequency corresponds to the highest frequency the basic version of the TV scope can be expected to display. At the end of the chain (at point 5) a clean, time-expanded replica of the original analogue input signal is now available for display purposes. Or rather, not quite. The A-channel output is amplified slightly (approximately times three) and a DC component is added to
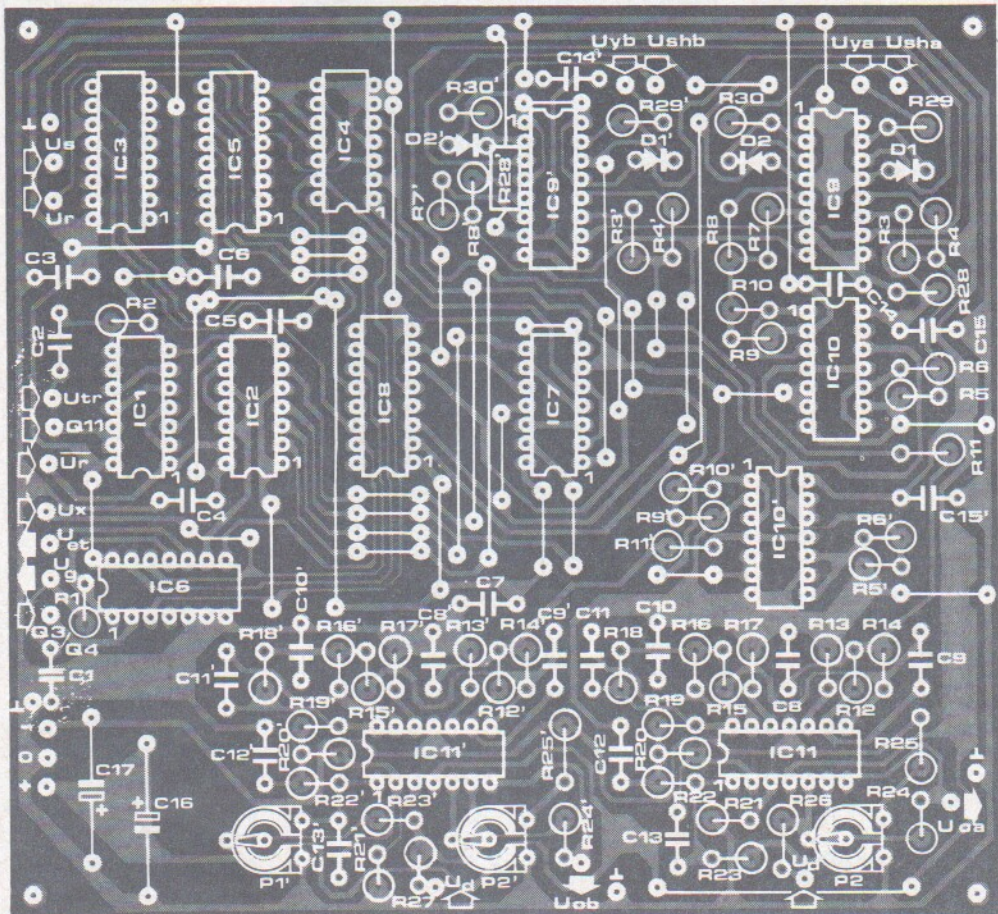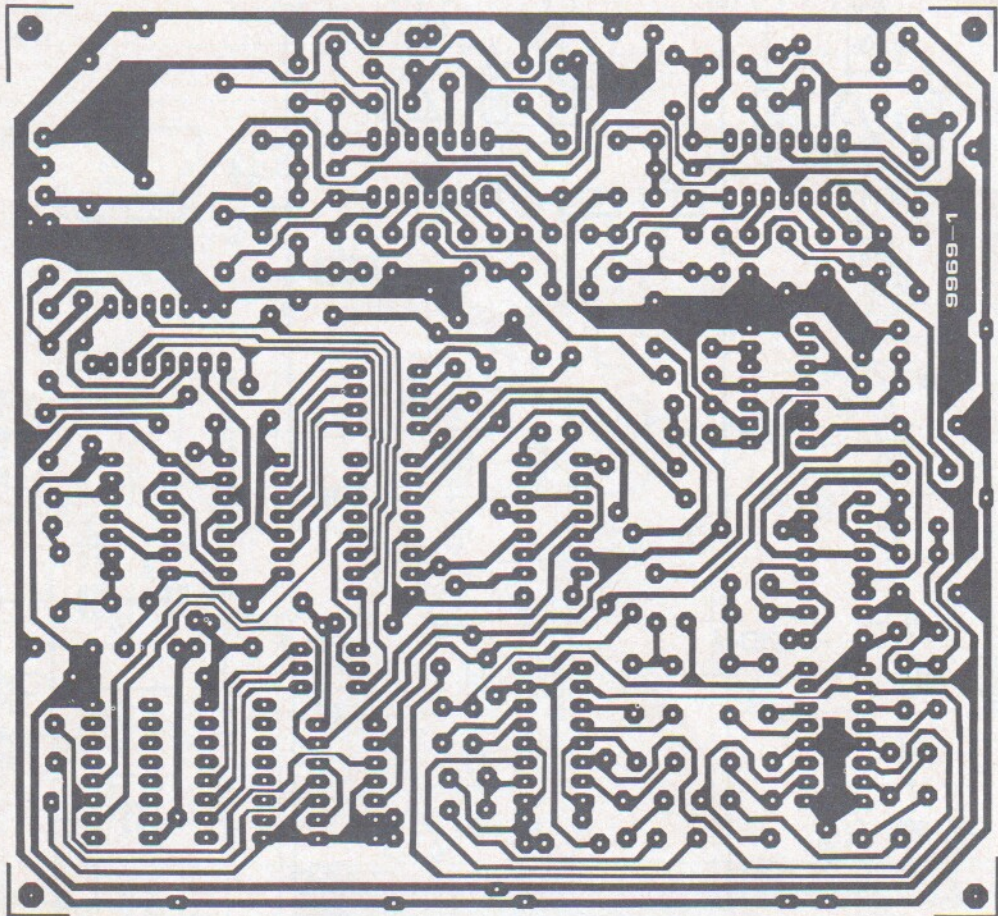
8b

A1 . . . A4 = IC11 = LM 324
A1' . . . A4' = IC11' = LM 324
D1 . . . D2 = DUS
D1' . . . D2' = DUS

79025 8b

9
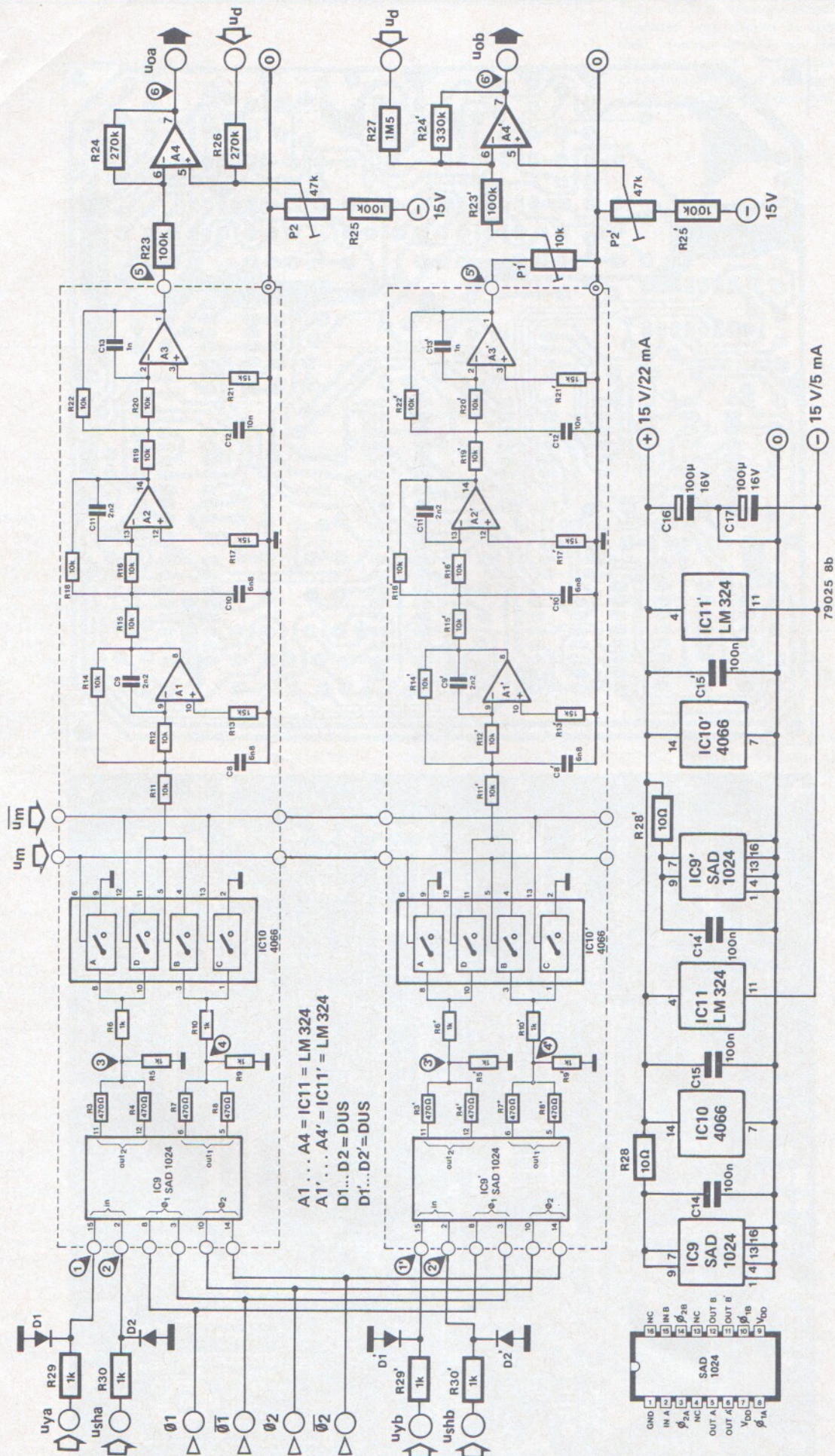
9969-1

Table 4

Supply currents (mA)

circuit	board no.	+15 V	−15 V
input amplifier*	9968-1	5 . . . 9	5 . . . 7
main board	9968-2	45	20
video mixer	9968-3	9	—
memory board	9969-1	22	5
trigger circuit	9969-2	25	20 . . . 25
input timebase VHF/UHF	9969-3	20	—
TV modulator	9967	15	—
sync circuit**	9968-4	4	—

* two required for two-channel version of TV scope
** only required for basic version of TV scope

Figure 9. The 'memory board' (EPS 9969-1). Insulated wire should be used for the wire links wherever there is any danger of inadvertent short circuits. Furthermore, the resistors are mounted 'on end' in order to keep the size of the board within reasonable limits.

Table 5

Calibration data for the input timebase (see text)

time/div.	preset	calibration signal	IC5 (main board) pin no.	frequency	period time	number of divisions	f_sample
2 ms	P7	Q8	13	480 Hz	2.08 ms	1.0	31.98 kHz
1 ms	P6	Q7	4	960 Hz	1.04 ms	1.0	63.96 kHz
500 µs	P5	Q6	2	1.92 kHz	520 µs	1.0	127.9 kHz
250 µs	P4	Q5	3	3.84 kHz	260 µs	1.0	255.8 kHz
100 µs	P3	Q5	3	3.84 kHz	260 µs	2.6	639.6 kHz
40 µs	P2	Q3	6	15.38 kHz	65 µs	1.6	1599 kHz

Parts list memory board

Resistors:
R1 = 33 k
R2 = 22 k
R3,R3',R4,R4',R7,R7', R8,R8' = 470 Ω
R5,R5',R6,R6',R9,R9', R10,R10',R29,R29', R30,R30' = 1 k
R11,R11',R12,R12', R14 . . . R16,R14' . . . R16', R18 . . . R20,R18' . . . R20', R22,R22' = 10 k
R13,R13',R17,R17', R21,R21' = 15 k
R23,R23',R25,R25' = 100 k
R24,R26 = 270 k
R24' = 330 k
R27 = 1M5
R28,R28' = 10 Ω
P1' = 10 k preset
P2,P2' = 47 k (50 k) preset

Capacitors:
C1,C12,C12' = 10 n
C2 = 47 p
C3,C4,C6 = 22 n
C5,C7,C14,C14', C15,C15' = 100 n
C8,C8',C10,C10' = 6n8
C9,C9',C11,C11' = 2n2
C13,C13' = 1 n
C16,C17 = 100 µ/16 V

Semiconductors:
D1,D1',D2,D2' = DUS
IC1,IC2,IC7 = CD 4011
IC3,IC5 = CD 4020
IC4 = CD 4013
IC6 = CD 4081
IC8 = CD 4049
IC9,IC9' = SAD 1024 (Reticon)
IC10,IC10' = CD 4066
IC11,IC11' = LM 324

Sundries:
S1 = single-pole, single-throw switch

produce the final output, u_{oa}, which is fed to the input of the main board. The output of the B-channel undergoes similar treatment, with one minor difference. The fixed gain is slightly greater (R24' is slightly larger than R24), but this output level can be reduced by means of P1'. In effect, therefore, the total gain at the output of the B-channel can be adjusted to compensate for any minor differences in the attenuation caused by all the preceding stages. The adjustable DC shift set by P2 brings the output level into line with the sawtooth reference voltage on the main board.

One last point remains to be discussed: the 'trace distance' control, P2 in figure 4, which was mentioned briefly in connection with the board for the trigger circuit. The DC output from this potentiometer, u_d, is connected via R26 and R27 to the non-inverting input of the A-channel output amplifier and the inverting input of its partner in the B-channel. The term 'trace distance' aptly describes the function of this control: it shifts the two traces in opposite directions on the screen. At one end of the control range, the two traces will overlap (useful for comparison of minor differences between the two), whereas at the other end of the range the two traces are each approximately centered on their own half of the screen.

All the circuits shown in figures 8a and 8b are mounted on a single p.c. board, the memory board, as given in figure 9. In order to pack such a large number of components on a relatively small, single-sided board, the ease of construction has had to be sacrificed to some extent. Insulated wire will be required for

several of the wire links, to avoid the possibility of highly undesirable short circuits. Furthermore, in marked contrast to most Elektor boards, the resistors are mounted 'on end' and care should be taken to avoid shorts between the long connections to adjacent resistors. Switch S1 ('x magn') is mounted off board, as shown in figure 10.

The components for the B-channel normally have the same values as their opposite numbers in the A-channel (e.g. R3 = R3' = 470 Ω). There are two exceptions to this rule: R24 (A-channel) = 270 k, whereas R24' (B-channel) = 330 k; furthermore, preset P1' (B-channel) is not included in the A-channel circuit. If a single-channel version of the TV scope is required, all components for the B-channel may be omitted.
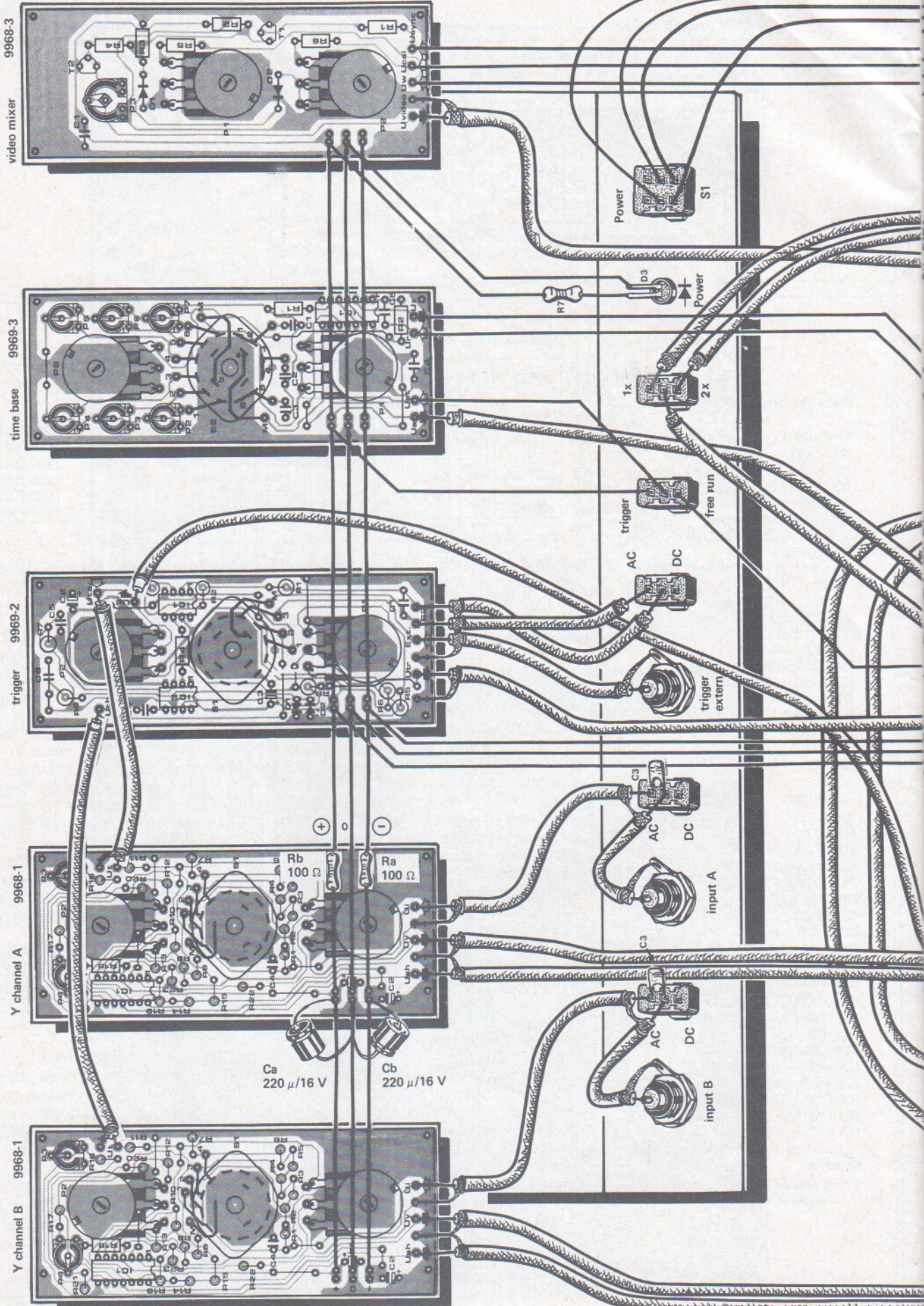
For test purposes, some voltages at the various test points are listed in Table 1. Further details are given in the description of the calibration procedure.

All the bits . . .

The three boards described in this article, together with those already described for the basic version, are all that is needed for the extended version of the TV scope. As an aid in procuring the necessary components, all the relevant parts lists have been compressed into one complete 'bulk parts list': 146 resistors, 10 potentiometers, 20 preset potentiometers, 82 capacitors, 34 ICs, 24 other semiconductors, 11 switches and a few 'sundries'.

For each component value or type, the quantity required for the complete extended version of the scope (basic version plus extension circuits) is listed
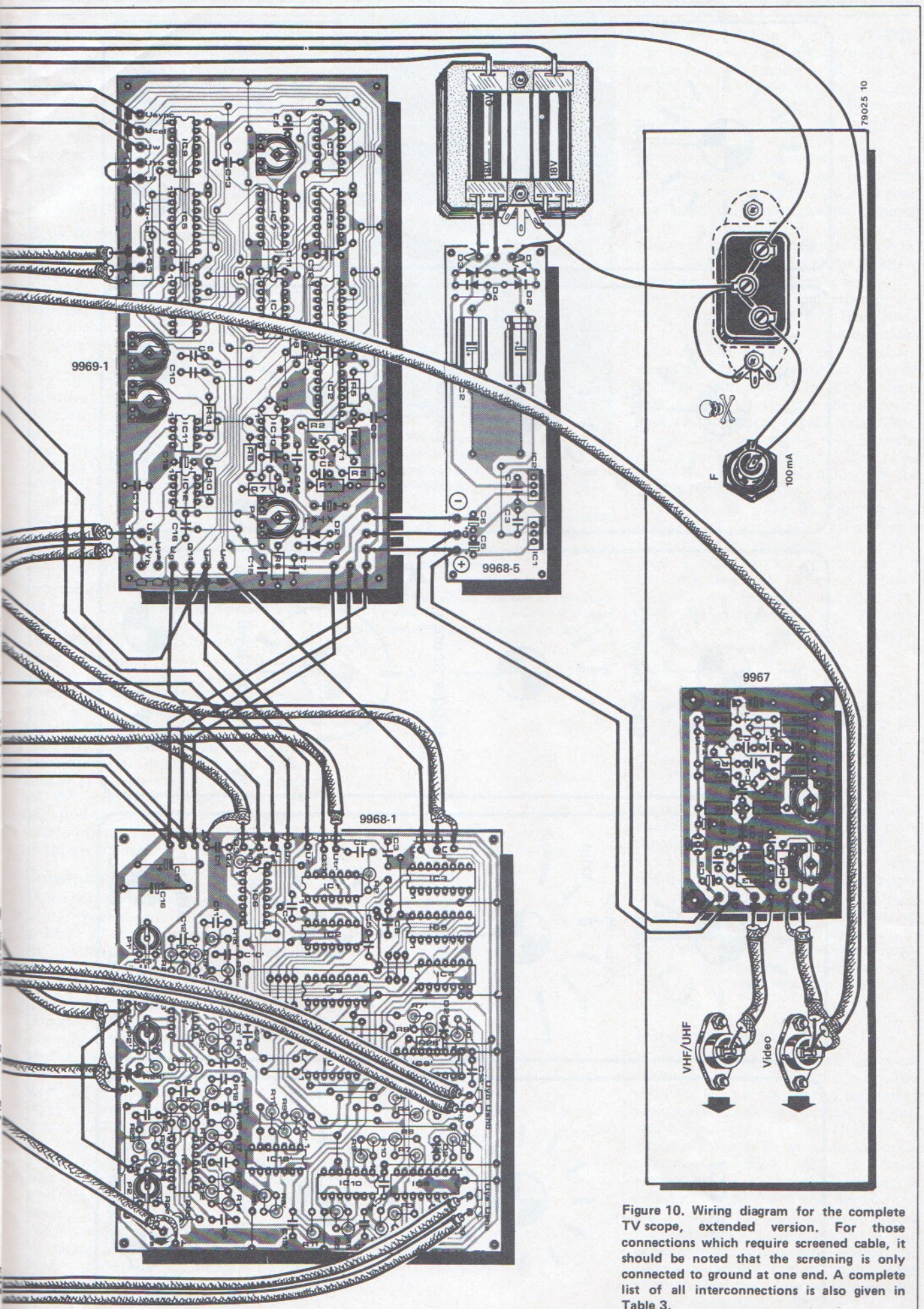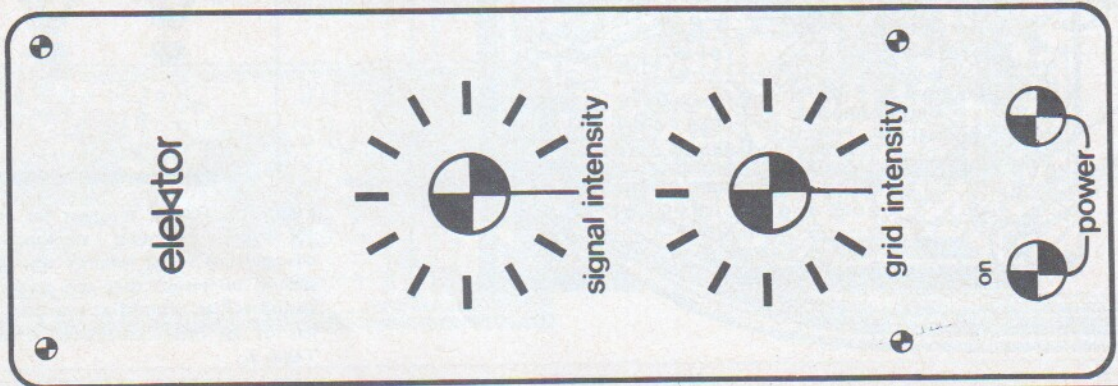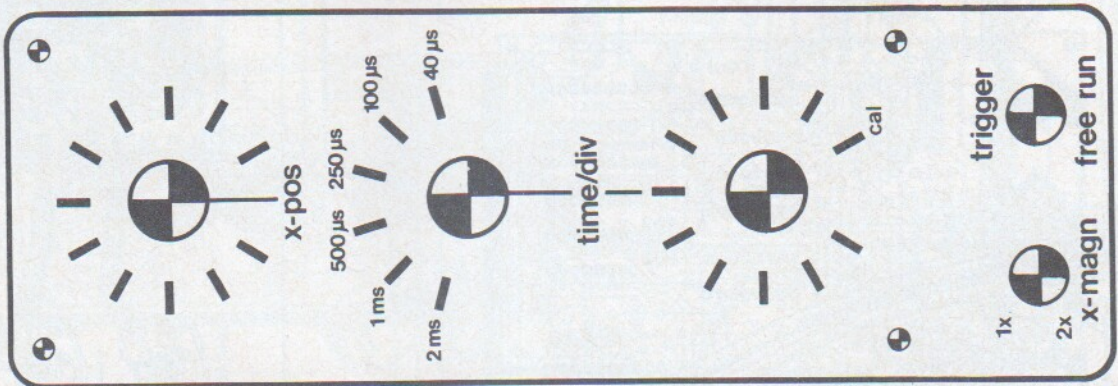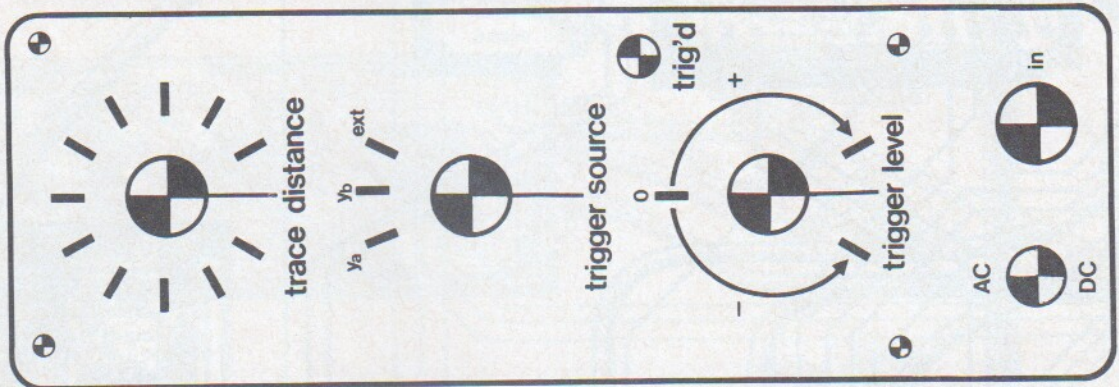
10

video mixer   9968-3

Uvideo Ūw Ucal Ūsync

Power   S1

R7   D3   Power

time base   9969-3

1x   2x

AC   DC   trigger   free run

trigger   9969-2

trigger extern

AC   DC   C3

Rb 100 Ω   Ra 100 Ω

⊕   0   ⊖

Y channel A   9968-1

input A

C3   AC   DC

Ca 220 μ/16 V   Cb 220 μ/16 V

AC   DC   input B

Y channel B   9968-1

9969-1

79025 10

F   100 mA

9968-5

9967

9968-1

VHF/UHF

Video

Figure 10. Wiring diagram for the complete TV scope, extended version. For those connections which require screened cable, it should be noted that the screening is only connected to ground at one end. A complete list of all interconnections is also given in Table 3.

11

Panel 1

y-pos

volts/div

10mV
100mV
1V
10V

cal

y
in
AC
DC

Panel 2

y-pos

volts/div

10mV
100mV
1V
10V

cal

y
in
AC
DC

Panel 3

trace distance

ya
yb
ext

trigger source

trig'd

trigger level

0
+
−

in
AC
DC

Panel 4

x-pos

time/div

40 µs
100 µs
250 µs
500 µs
1 ms
2 ms

cal

trigger

free run

1x
2x
x-magn

Panel 5

elektor

signal intensity

grid intensity

power

on
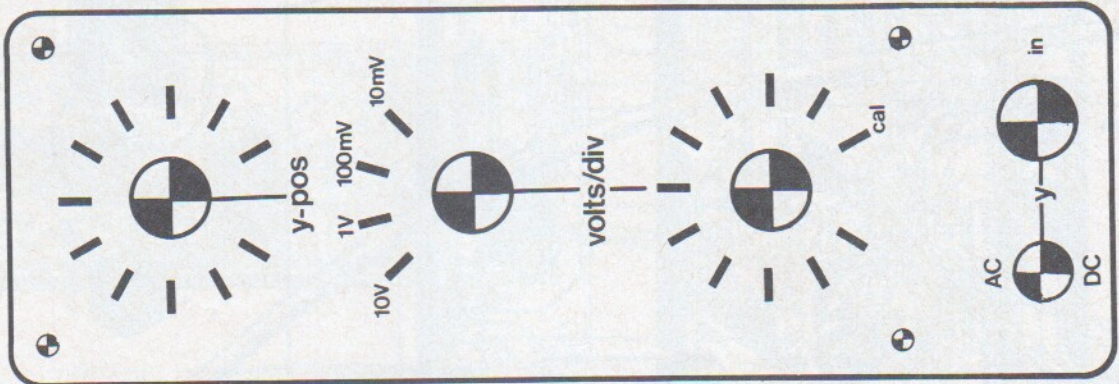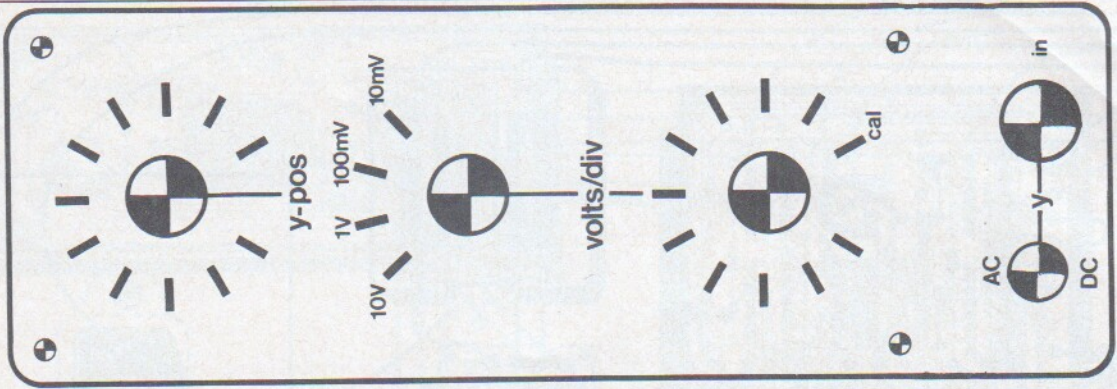
in one column; a second column gives the quantities required for extending the basic scope. Both columns are valid for the complete two-channel TV scope. To avoid misunderstandings: in the first column ('complete'), the components required for the UHF/VHF TV modulator are also listed (see the recent October issue, page 10-27). This was not the case in the bulk parts list for the basic version of the TV scope given in the October issue.

Back to the basic version

During extensive testing of the extended version of the TV scope, it was discovered that some samples of the analogue delay-line ICs exhibited noticeable anaomalies in their signal-storage characteristics. This effect is less apparent if larger signal levels are used.

For this reason, one or two minor modifications of the basic version are required if it is to be used in conjunction with the extension circuits. Note that there is no advantage in incorporating these modifications in the basic scope if it is not to be extended — although they won't do any harm, either.

As shown in Table 2, three resistor values are changed: two in each y-input amplifier (four resistors in total for a two-channel scope) and one resistor on the main board. These components are included in the bulk parts list.

The effect of the modifications in the input amplifier is to significantly increase its output level, making it more suitable for processing in the analogue delay lines. A minor disadvantage is a slight reduction in bandwidth: − 6 dB at 100 kHz, instead of the original − 3 dB. Since the input signal to the main board is now at a higher level, the amplitude of the reference sawtooth must be increased in step. This is accomplished by modifying the value of one resistor on the main board.

Construction

The connections between the various boards are shown in figure 10. The position of the boards corresponds to the lay-out of the front panel shown in figure 11. The front-panel controls are listed in Table 6, with a brief description of their function.

Construction of the complete unit is basically similar to that of the basic version, as described in the October issue. Suitable cases are available from West Hyde (type DES 306) and GSA (type V7004). As with the basic version, it is advisable to screen the more sensitive circuits (input amplifiers, trigger circuit, input timebase and video mixer) with strips of copper laminate board which can be soldered to the earth plane.

The interconnections between the various boards are also listed in Table 3. For those connections which require screened cable, it should be noted that the screening is only connected to supply common at one end, to avoid ground loops.
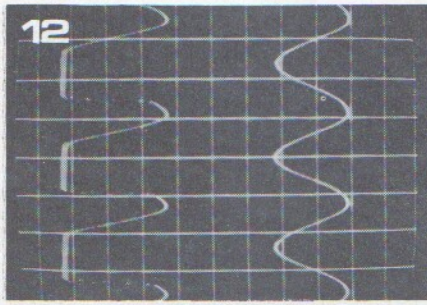
Figure 11. Front panel layout for the extended version of the TV scope. Table 6 lists the front-panel controls, with a brief description of their function.

Figure 12. A word of warning! The bucket-brigade delay-lines have a limited dynamic range, and excessively large input signals will be clipped, as shown here.

Figure 13. Detail of the interior of the prototype. To the left of the picture, the main supply regulators are mounted on the back panel.

As shown in figure 10, the supplies to the input amplifiers are decoupled by means of two resistors, R_a and R_b (100 Ω), and two electrolytic capacitors (C_a and C_b, 220 μ/16 V). These components are mounted 'off board', as shown.

When comparing the basic version with the extended version, it may be noticed that the 'reset' output is omitted from the latter. This provision is not required in the extended version, since it has adequate triggering facilities. However, if the 'reset' output is already mounted there is no point in removing it.

The two IC voltage regulators in the main supply should be adequately cooled, for instance by mounting them (with mica insulation!) on the back of the case. The main reason for this is to keep the heat out of the case: experiments have shown that the performance of the bucket-brigade delay lines, in particular, deteriorates rapidly as the ambient temperature rises. There is no need to cool the regulator on the UHF/VHF modulator board.

A further aid when constructing and testing the TV scope is the list of supply currents for the various circuits, given in Table 4. The current consumption of the complete TV scope (extended version, two-channel, including the VHF/UHF modulator) is approximately 150 mA from the positive supply (+15 V) and 55 mA from the − 15 V supply.

Before applying power for the first time, it is advisable to first give the wiring a final 'once-over', checking for inadvertent short circuits. Some photos of the completed unit are given in figures 13 . . . 15. To play things safe, the supply can be connected to the various boards one at a time: first disconnect the output from the main supply, apply power and check the +15 V and −15 V outputs. The other boards are then connected to the supply one at a time, checking the supply current against Table 4. Differences of up to 10% are no cause for alarm.

Once all the boards have been connected up and checked in this way, it is time for the final step: the calibration procedure.

Initial calibration

In spite of the 20 presets involved, calibration of the TV scope is not too difficult. The only measuring equipment required is a standard universal meter, with a sensitivity of at least 10 kΩ/V.

The first step is to calibrate the components associated with the basic version of the TV scope. The relevant adjustment procedures were discussed in the article in the October issue. If one has already built and calibrated the basic version and is now adding the extension circuits, the only adjustment affected by the component modifications according to Table 2 is the setting of P1 on the main board. How-
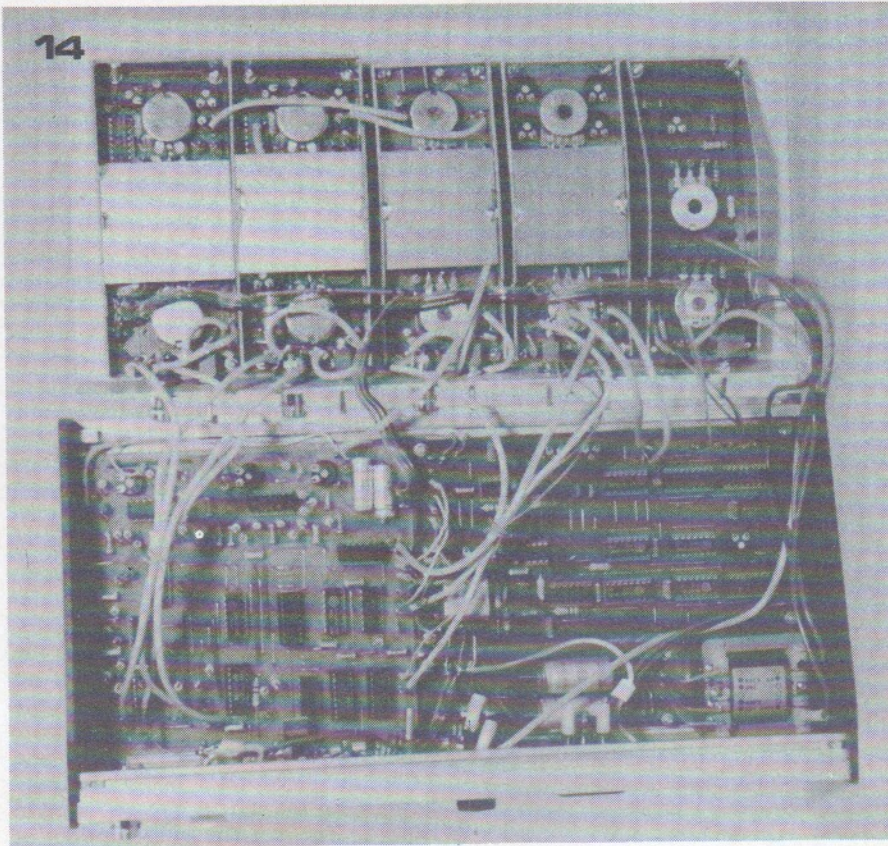
Figure 14. Complete unit, with all wiring and screening panels in place.

Figure 15. A 10 kHz square-wave on an ancient TV set and a 20 kHz sine-wave on a more recent model . . .

ever, final adjustment of this preset is part of the calibration procedure about to be described. The setting of all other presets in the basic version remains unaltered, with the exception of P3 and P4 on the input amplifier board(s). These two potentiometers have no influence on the operation of the basic version.

Having completed the adjustment procedure outlined for the basic version, the next step is to adjust P4 in the input amplifier. For this initial (approximate) adjustment, no input signal is required. The controls of the y-amplifier are set as follows: 'volts/div' selector switch a 10 V and corresponding potentiometer fully clockwise ('cal'); 'AC/DC' switch in position 'AC'. Initially, set P3 and P4 in their mid-positions. Measure the voltage at test point 1 (or 1') on the memory board, and adjust the 'y-pos' control to obtain 5.5 V at this point. Now measure the voltage at point 2 (2') and adjust P4 until 5.5 V is obtained at this point also. This adjustment is sufficiently accurate for the time being. Operation of the bucket-brigade memories can now be checked. To this end, the 'trace distance' control is turned fully anti-clockwise; the 'time/div' switch is set to 500 μs, and the corresponding potentiometer fully clockwise ('cal'); the 'trigger/free run' switch is set to 'free run'; all presets on the input timebase board are set to their mid-position. The voltages at test points 3 and 4 (3' and 4') should now be approximately 3 V; furthermore, it should be possible to vary these voltages over a total range of about 1 V (e.g. from 2.5 V to 3.5 V) by means of the 'y-pos' control. If this is the case, the memories

are in working order.

The next point to check is the voltage at point 5 (5'). This should be approximately equal in value to the voltage at points 3 and 4 (3' and 4'), but negative with respect to supply common. In other words, if the voltages at points 3 and 4 are indeed 3 V, the voltage at point 5 should be approximately −3 V. This being the case, the next step is to measure the voltage at point 6 (6') and set it to approximately +4 V by means of P2 (P2'). For the time being, P1' is set to the mid-position.

The output of the TV scope should now be connected to a TV receiver. P1 on the main board (basic version) is adjusted until vertical lines appear that are not part of the calibration graticule. In principle, there should be four of these lines, but it is possible that one or more of them are actually off the screen. If this is the case, the missing lines can be brought into the picture by adjusting P2 (P2') on the memory board.

One pair of lines is generated by the A-channel. This pair can be found by operating the corresponding 'y-pos' control: the lines can be shifted over a total range of approximately 4 scale divisions, or just less than half the width of the screen (the range is limited by the maximum output swing of the bucket-brigade memories). The y-position controls for both channels are now set so that the corresponding vertical lines are mid-way between their two extreme positions. Finally, P4 on the input amplifier board(s) is adjusted until the corresponding pair of lines become one: the zero-volt line for that particular channel.

This completes the initial setting-up

procedure; the next step is the final calibration.

Final calibration

If a dual-channel version of the scope is being calibrated, it is a good idea at this point to adjust P2' on the memory board so that the zero-volts line for the B-channel is moved to the edge of the screen. For the present, the A-channel display is the only one which is of interest.

The first step is to calibrate the input timebase. For this, a stable and accurate AC test signal is required. Fortunately, no expensive test equipment is required: a suitable signal can be derived from the TV scope itself. The input of the y_a input amplifier is connected to pin 2 ('Q_6') of IC5 on the main board (basic version). Since this is a digital IC and the relevant oscillator is crystal-controlled, the signal at this point is known: it is a squarewave with a peak-to-peak amplitude of 15 V and a frequency of 1.92 kHz, corresponding to a period time of 520 μs.

With the 'trigger source' selector switch set to 'y_a', it should now be possible to adjust the 'trigger level' control so that the 'trig'd' LED lights. This adjustment is fairly critical. The 'trigger/free run' switch is now set in position 'trigger', whereupon a stable picture should appear on the screen. The square-wave trace can be moved horizontally and vertically across the screen by means of the 'y-pos' and 'x-pos' controls, respectively. Note that the y-position control should be adjusted so that the trace is centered in its range, as otherwise the signal will be clipped by the delay-lines. At this stage, the display will almost

Table 6

Front-panel controls

function	part no.	figure no.*	remarks
power	S1	B14	mains switch
signal intensity	P1	B7	intensity of trace
grid intensity	P2	B7	intensity of calibration graticule
x-pos	P8	E6	trace position on x-axis
time/div**	S2,P1	E6	x-axis scale
x-magn	S1	E8a	'1x': scale corresponds to that selected by 'time/div' switch; '2x': x-axis scale multiplied by 2
trigger/free run	S1	E6	trigger circuit enable; select 'free run' in absence of trigger signal
trigger source	S1	E4	selector switch for trigger signal
trig'd	D2 (LED)	E4	indicator LED, lights if trigger signal is present
trigger level	P1	E4	reference level for signal selected as 'trigger source'
trace distance	P2	E4	sets distance between the two traces for A- and B-channel
y-pos	P2	B2	trace position on y-axis
volts/div**	S1,P1	B2	y-axis scale
AC/DC	S2	B2/E4	AC or DC coupling of input signal at respective input

* Figure numbers preceded by the letter 'B' refer to figures in the article 'TV scope, basic version' in the October issue; figure numbers preceded by letter 'E' refer to figures in this article.
** The values given for the various switch positions are only valid if the corresponding potentiometer is in position 'cal'.

certainly consist of two square-waves, one corresponding to each of the delay-lines in the A-channel. The differences between the two signal paths can now be compensated, by means of presets P3 and P4 in the A- input amplifier. The slight difference in gain between the two delay-lines can be compensated for by adjusting P3 until the amplitude of the two square-wave signals is identical; the varying DC shifts are brought into line by setting P4 so that both traces are symmetrical around the same centre-line — in other words, the y-position of both traces should be identical. Since these adjustments interact, this adjustment procedure must be repeated until no further improvement is obtained.

Having reduced the display to a single square-wave signal, the next step is to calibrate the input timebase. With the 'time/div' switch in position 500 μs and the corresponding potentiometer in position 'cal', one complete period of the square-wave should correspond to a fraction over one division — 520 μs is 1.04 divisions. The preset potentiometer P5 on the input timebase board is adjusted until the correct period length is obtained.

The 'volts/div' selector switch should still be set at '10 V'. This means that the 15 V peak-to-peak square-wave should correspond to an amplitude of 1.5 divisions. This result can be obtained by (re-) adjusting P1 on the main board of the basic version. It will be noticed that this adjustment also influences the y-position, but this shift can be compensated for by readjusting P2 on the memory board.

The same adjustment procedure can now be carried out for the second (B-) channel, after first bringing the corresponding 'zero-volts line' back into the centre of the picture by re-adjusting P2' on the memory board. There is, however, one difference in the calibration procedure: the amplitude setting (for 1.5 horizontal divisions) is carried out by means of P1' on the memory board, instead of using P1 on the main board.

For the B-channel, the procedure can therefore be summed up briefly as follows: P3 and P4 on the B-channel input amplifier are adjusted until the two square-wave signals overlap; the timebase, obviously, needs no re-adjustment, since it is common to both channels; P1' on the memory board is so that the amplitude of the square-wave signal corresponds to 1.5 horizontal divisions; finally, P2' on the memory board is adjusted to compensate for DC shift (y-position).

Final adjustment of P2 and P2' is now in order. The 'trace distance' control should still be set fully anti-clockwise ('overlap') and the 'y-position' controls are both set so that the corresponding traces (with no input signal applied) are midway between their two extreme positions. P2 and P2' can now be adjusted so that the corresponding traces overlap in the exact centre of the screen.

The remaining presets in the input timebase now require attention — P5 has already been adjusted. The other presets are calibrated in the same way — the 'trigger level' control being manipulated to obtain a stationary display. Suitable calibration signals are derived from the TV scope itself, and the relevant preset on the input timebase board is adjusted until the correct period length is ob-

tained. All timing signals are derived from IC5 on the main board. All relevant details are given in Table 5: the position of the 'time/div' selector switch; the corresponding preset potentiometer; the necessary calibration signal, and the pin number (of IC5) from which this signal can be obtained; the frequency and period time of this signal; the number of divisions that corresponds with one complete period of this signal when the preset is correctly adjusted. The last column in this table gives the correct frequency (f_{sample}) of the u_{sample} signal, for the benefit of those who have access to a frequency counter: setting the 'trigger/free run' switch in position 'free run' causes the timebase oscillator to run continuously, so that it can be measured and calibrated accurately without the need for recourse to calibration signals. For completeness' sake, the calibration data for P5 are also listed in Table 5, even though this adjustment has already been completed at an earlier stage.

A possible complication in the '2 ms/div' position is that it may prove impossible to obtain a stationary display. The reason for this problem is that in this case the 'read' cycle may take so long that it overlaps the 'display' cycle — the memories are read out before they have been completely read in! In this case, triggering is not possible and the 'trigger/free run' switch may as well be set in the 'free run' position.

The adjustment of the input timebase was the last step in the calibration procedure. It may, however, prove useful to give all adjustments a final once-over. In particular, presets P3 and P4 in the input amplifiers may require

Bulk parts list for extended version of TV scope
(two input amplifiers, main board, video mixer, input timebase, trigger circuit, memory board, VHF/UHF TV modulator and supply — note that the sync circuit is not required for the extend version)
'complete' (second column) gives the total number of each component type required for the TV scope in extended version; 'extension only' (third column) lists the components required to extend the basic version

Resistors:

value	complete	extension only
10 Ω	3	2
47 Ω	4	1
68 Ω	1	
100 Ω	4	2
150 Ω	1	
220 Ω	2	1
270 Ω	1	
330 Ω	1	1
470 Ω	10	8
820 Ω	2	
1 k	21	12
1k5	1	
2k2	1	
3k3	1	
4k7	1	1
5k6	4	2
6k8	1	
8k2	4	
10 k	26	20
12 k	2	
15 k	12	6
18 k	2	
22 k	3	2
33 k	3	1
47 k	2	
82 k	6	
100 k	15	7
220 k	2	
270 k	2	2
330 k	1	1
390 k	1	1
680 k	1	1
820 k	2	
1M5	1	1
2M2	2	2

Linear potentiometers:

value	complete	extension only
2k2 (2k5)	1	1
10 k	2	
22 k (25 k)	1	1
47 k (50 k)	1	1
100 k	3	1
220 k (250 k)	1	
470 k (500 k)	1	

Preset potentiometers:

value	complete	extension only
1 k	2	
2k2 (2k5)	2	
4k7 (5 k)	2	
10 k	4	1
47 k (50 k)	10	8

Capacitors:

value	complete	extension only
3p3	1	1
8p2	3	
15 p	2	
18 p	1	1
22 p	1	
33 p	2	
47 p	1	1
82 p	1	
100 p	5	2
120 p	1	
220 p	1	
560 p	1	1
1 n	2	2
2n2	4	4
6n8	4	4
10 n	5	3
22 n	4	4
100 n	24	8
1 μ	1	1
1 μ/25 V tantalum	10	2
2μ2/25 V tantalum	1	
4μ7/35 V tantalum	1	
100 μ/16 V	2	2
220 μ/16 V	2	2
470 μ/35 V	2	

Semiconductors:

type	complete	extension only
CD 4011	6	2 (3)*
CD 4012	1	
CD 4013	2	1
CD 4017	1	
CD 4020	2	2
CD 4040	1	
CD 4049	1	1
CD 4066	2	2
CD 4068	1	
CD 4071	1	
CD 4081	1	1
CD 4093	1	1
709	3	1
741	2	1
LM 324	2	2
TL 084	2	
SAD 1024	2	2
78L05	1	
7815	1	
7915	1	
TUN	3	
TUP	1	
BF 494	3	
BFY 90	1	
DUS	9	5
1N4001	4	
1N4148	1	
LED	2	1

switches:

type	complete	extension only
single-pole, single-throw	4	2
double-pole, single-throw	1	
double-pole mains switch	1	
single-pole, 3-way	2	2
single-pole, 4-way	2	
2-pole, 6-way	1	1

sundries (complete):

	complete	extension only
4.433 MHz (colour TV) crystal	1	1
27 MHz (approx.) crystal	1	1
1 μH mini-choke	1	1
100 mA fuse	1	1
2x18 V/250 mA mains transformer	1	1

* 2 off if the CD 4011 originally mounted on the sync board can be re-used on the memory board, otherwise 3 off.

minor re-adjustment after the 'bucket-brigade' ICs have warmed up. These chips have proved to be rather sensitive to changes in ambient temperature, with shifts in output level that vary from one sample to the next. This may result in trace 'flicker' after the unit has warmed up (actually, two separate traces are displayed very close together, giving a flickering effect) but this can be eliminated by slight re-adjustment of P3 and P4. A good test signal for this final adjustment is a sine-wave with an amplitude of about three divisions on the screen.

'There's a hole in the bucket'

The Reticon SAD1024 bucket-brigade delay-line is a useful IC, but it is intended primarily for audio reverberation systems. Tests have shown that a small percentage of these ICs contain one or more 'leaky buckets'. Since, in the TV scope, the input signal is stored for a short time without being shifted, one or more of the input samples may remain in a leaky bucket for a relatively long period. This results in noticeable 'dips' in the trace. For this reason, the signal level has been boosted (by means of the modifications listed in Table 2). The dips are now reduced, relative to the signal, to the point were they are no longer a nuisance.

For the same reason it is advisable, if triggering is not required, to select the 'free run' position (e.g. when measuring DC voltages). Since the timebase then runs continuously, the contents of the memories are refreshed right up to the point where the display cycle starts, giving the best results.

Litterature:
Elektor, May 1978, P.5-20: 'TV scope using bucket-brigade memory';
Elektor, October 1978:
– page 10-03: 'an introduction to the TV scope';
– page 10-30: 'TV scope, basic version';
– page 10-27: 'VHF/UHF TV modulator';
– page 10-44: 'analogue reverberation unit';
Elektor, November 1978, page 11-25: 'extending the TV scope'.

extending the elektor piano

Although the compass of the Elektor piano which was published in September of this year (Elektor No. 41) originally covered only five octaves, it was stated then that thanks to its modular design, it could be easily extended to span 6, 7 or even 8 octaves. Judging by the reaction from our readers, it does in fact appear that there is a considerable interest in an electronic piano with a full 8-octave compass, i.e. ninety-six keys and a fundamental frequency range extending from 17.4 to 4148 Hz. For this reason the following short article provides readers with details of the changes in component values and of the additions to the filter circuits which will be required to extend the compass of the Elektor piano.

1
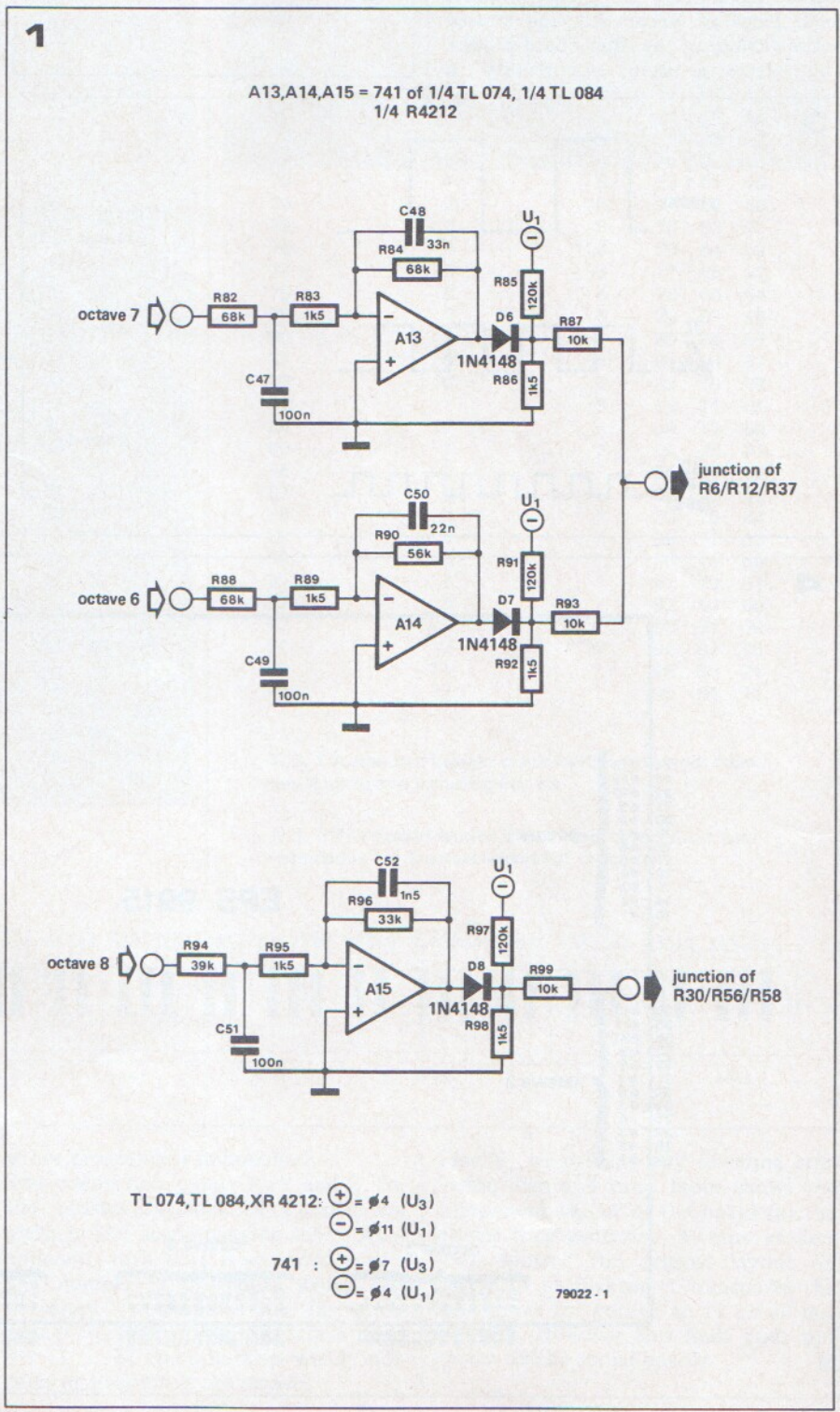
Firstly there is the question of the values of the discharge resistors R1 through R12 on the extra octave boards (EPS 9981). For the highest octave, i.e. octave 8, R1 through R6 = 100 k, and R7 through R12 = 120 k. For octave 6, the second *lowest* octave, R1 through R6 = 1 M, and R7 through R12 = 1M2, whilst for octave 7, the lowest octave, R1 through R6 = 1M5, and R7 through R12 = 1M8.

The original board layout for the filter circuits (EPS 9981) was designed for the 5-octave version. Thus, depending on however many extra octaves are required, it will be necessary to incorporate one or more of the additional filter circuits shown in figure 1. These will, of course, have to be mounted on, e.g. a small piece of Veroboard, and housed in the vicinity of the filter board proper. This seems a useful opportunity to mention an interesting possibility for those readers who, without wishing to redesign the voicing circuits, might like to experiment with altering the sound of the original piano. The following tip has been received from a Dutch reader, a Mr. Hulshoff from Rotterdam, who recommends introducing even harmonics (the second, fourth, sixth harmonic etc. of the fundamental) into the squarewave outputs of the master tone generator, which being symmetrical, are presently composed only of the fundamental and a number of odd harmonics.
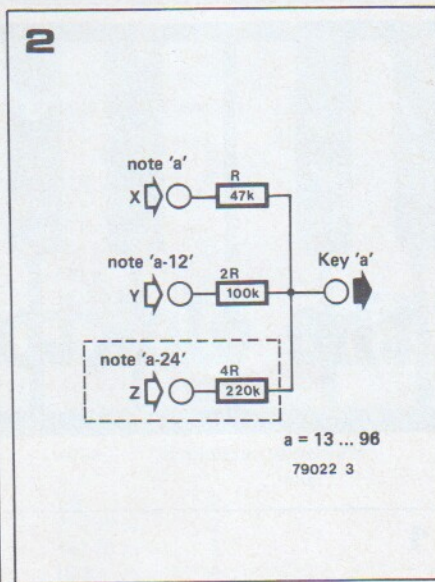
With the aid of the resistor network shown in figure 2, the master tone generator outputs are not fed straight to the corresponding inputs of the keying circuits; rather they are summed with one, or in most cases, two squarewave signals which successively have twice the frequency but half the amplitude of the

2

note 'a'
X ▷ ○ — [R 47k]
note 'a-12'
Y ▷ ○ — [2R 100k] — ○ Key 'a' ▶
note 'a-24'
Z ▷ ○ — [4R 220k]

a = 13 ... 96
79022 3

3

I
f
(100%)

II
2f
(50%)

III
4f
(25%)

I+II+III
(octaves 2 ... 7)

I+II
octave 1

79022 2

4

octave 8

octave 1

octave 2

EPS 9915

octave 7

octave 6

octave 3   octave 4   octave 5

79022 4
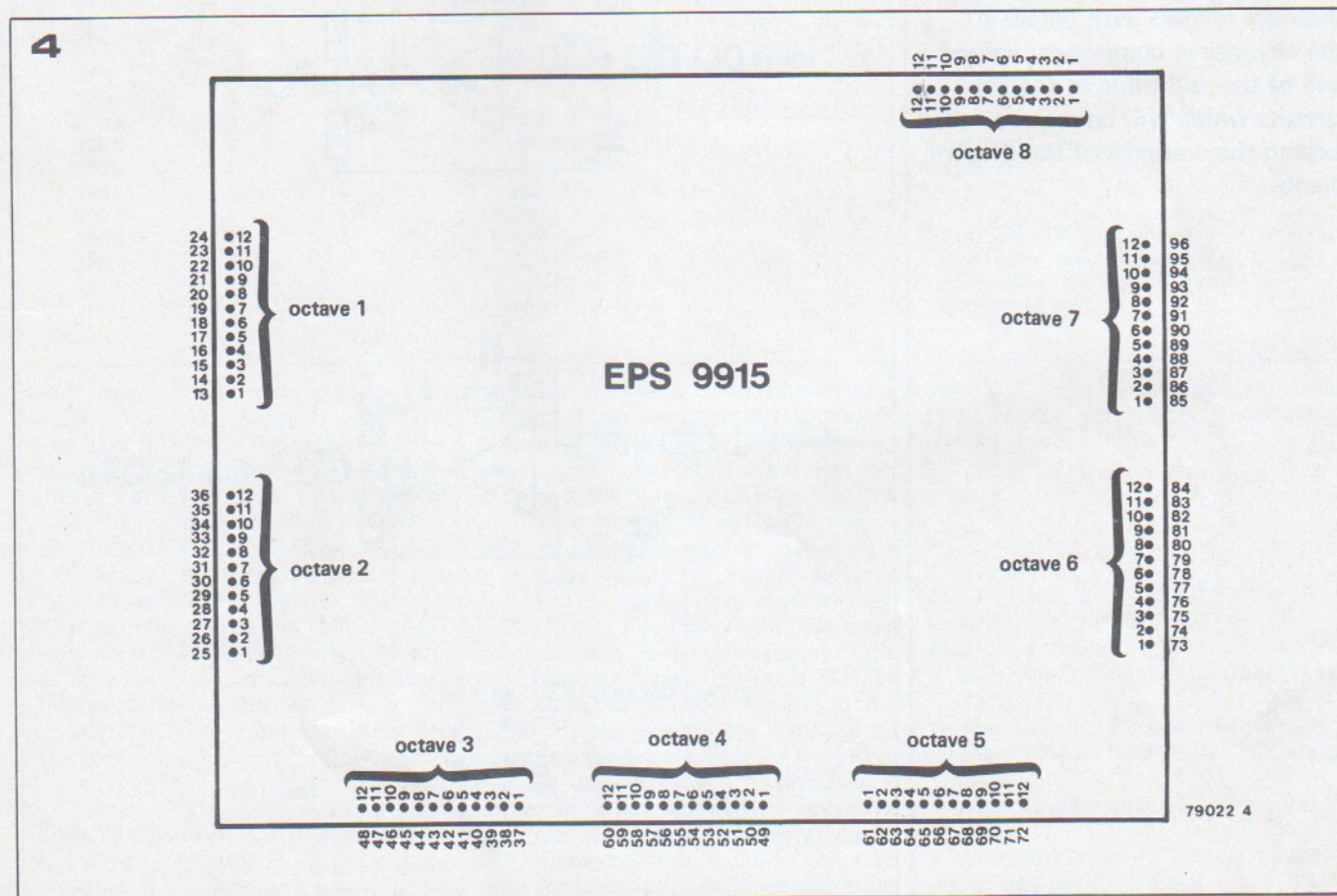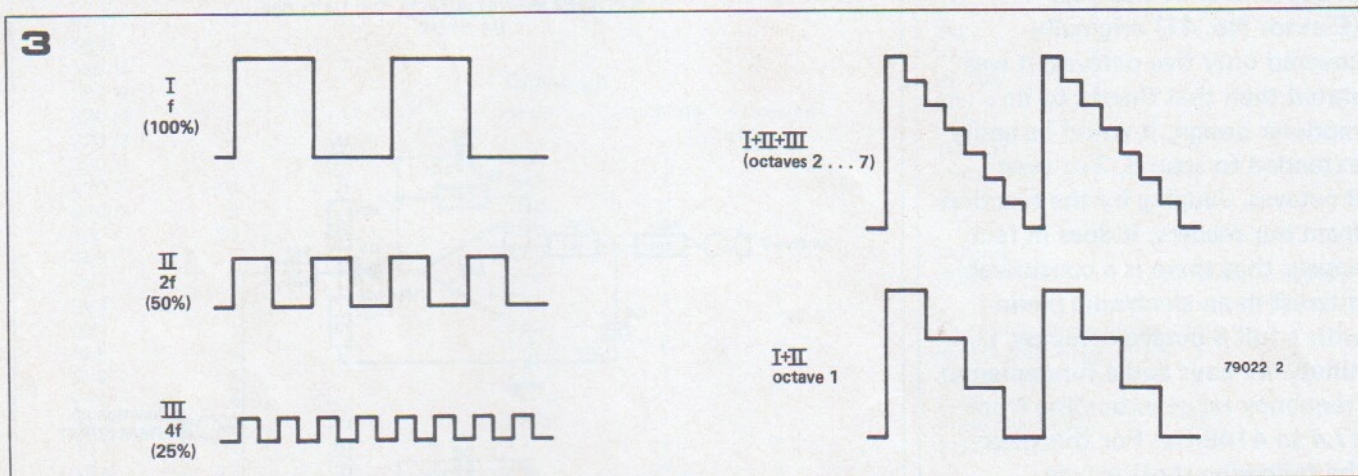
key number	number...	of octave...	X	Y	Z
1	1	8	—	—	—
2	2	8	—	—	—
3	3	8	—	—	—
4	4	8	—	—	—
5	5	8	—	—	—
6	6	8	—	—	—
7	7	8	—	—	—
8	8	8	—	—	—
9	9	8	—	—	—
10	10	8	—	—	—
11	11	8	—	—	—
12	12	8	—	—	—
13	1	1	13	1	—
14	2	1	14	2	—
15	3	1	15	3	—
16	4	1	16	4	—
17	5	1	17	5	—
18	6	1	18	6	—
19	7	1	19	7	—
20	8	1	20	8	—
21	9	1	21	9	—
22	10	1	22	10	—
23	11	1	23	11	—
24	12	1	24	12	—
25	1	2	25	13	1
26	2	2	26	14	2
27	3	2	27	15	3
28	4	2	28	16	4
29	5	2	29	17	5
30	6	2	30	18	6
31	7	2	31	19	7
32	8	2	32	20	8
33	9	2	33	21	9
34	10	2	34	22	10
35	11	2	35	23	11
36	12	2	36	24	12
37	1	3	37	25	13
38	2	3	38	26	14
39	3	3	39	27	15
40	4	3	40	28	16
41	5	3	41	29	17
42	6	3	42	30	18
43	7	3	43	31	19
44	8	3	44	32	20
45	9	3	45	33	21
46	10	3	46	34	22
47	11	3	47	35	23
48	12	3	48	36	24
49	1	4	49	37	25
50	2	4	50	38	26
51	3	4	51	39	27
52	4	4	52	40	28
53	5	4	53	41	29
54	6	4	54	42	30
55	7	4	55	43	31
56	8	4	56	44	32
57	9	4	57	45	33
58	10	4	58	46	34
59	11	4	59	47	35
60	12	4	60	48	36
61	1	5	61	49	37
62	2	5	62	50	38
63	3	5	63	51	39
64	4	5	64	52	40
65	5	5	65	53	41
66	6	5	66	54	42
67	7	5	67	55	43
68	8	5	68	56	44
69	9	5	69	57	45
70	10	5	70	58	46
71	11	5	71	59	47
72	12	5	72	60	48
73	1	6	73	61	49
74	2	6	74	62	50
75	3	6	75	63	51
76	4	6	76	64	52
77	5	6	77	65	53
78	6	6	78	66	54
79	7	6	79	67	55
80	8	6	80	68	56
81	9	6	81	69	57
82	10	6	82	70	58
83	11	6	83	71	59
84	12	6	84	72	60
85	1	7	85	73	61
86	2	7	86	74	62
87	3	7	87	75	63
88	4	7	88	76	64
89	5	7	89	77	65
90	6	7	90	78	66
91	7	7	91	79	67
92	8	7	92	80	68
93	9	7	93	81	69
94	10	7	94	82	70
95	11	7	95	83	71
96	12	7	96	84	72

N.B. 1 A standard upright piano keyboard extends from key 4 up to and including key 88.

N.B. 2 The shaded section of the table covers those keys included in the 5-octave version of the piano.
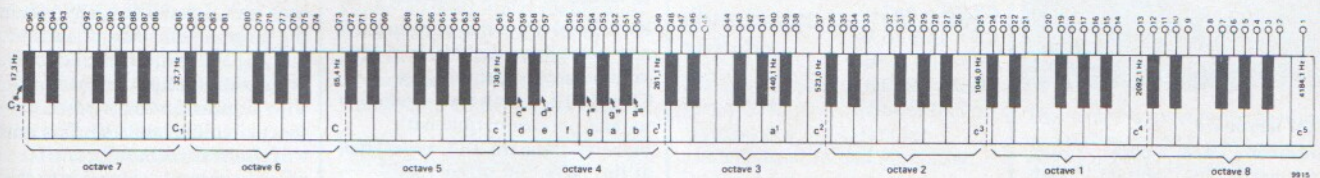
original squarewave. As figure 3 clearly shows, the result is a staircase waveform which has a high proportion of even harmonics. (Obviously, in the case of octave 1, i.e. the second highest octave, it is only possible to sum the corresponding master tone generator outputs with *one* signal of twice the frequency, i.e. that of octave 8 — the highest octave, whilst in the case of octave 8 itself, the above procedure is impossible).

The connection points X, Y and Z, for the resistor network of figure 2 are given in the accompanying table. As is apparent, each note is accorded both a key number (a procedure which was necessary to simplify the figure on page 9-10, Elektor 41), an octave number (1 . . . 8) and a number which indicates its position in the octave.

Finally, to prevent any lingering confusion figure 4 once more shows the component layout of the board for the master tone generator. As most readers soon spotted the original layout on page 9-11 of Elektor 41 transposed the indications for notes 1 to 12 on all but octave 4. However this fault does not appear on the actual boards.

missing link

Modifications to
Additions to
Improvements on
Corrections in
Circuits published in Elektor

Digital reverberation unit
Elektor 37, May 1978, p. 5-08. Three possible pin-compatible equivalents are listed for the 1024-bit shift register. However, when using the MM5058 the dissipation may in some cases be on the high side. The reason for this is the clock pulse level: 12 V top-top, whereas this particular IC is quite happy with only 5 V. A simple circuit modification (only required if the MM5058 is used!) is given below.
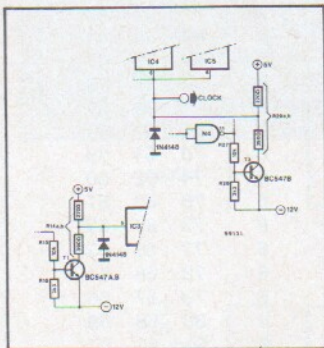
FET audio amplifier
Summer Circuits 1978, circuit no. 39. Components C5, C7, R14 and D9 are shown in the circuit as connected to supply common. However, they should be connected to the R7/R25 junction, as shown in the diagram below.
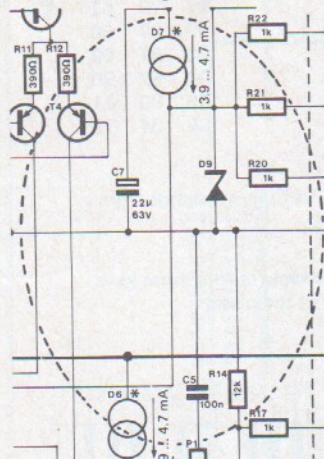
ASCII keyboard
November 1978, page 11-06. The parts list should read: R1 = 680 k; R2 = 100 k. The keyboard switches are listed as type MM9; the full type number is MM9-2. In Table 3, the last few lines are incorrect. The corrected section of the table is given below.

CTL + M = CR (CARRIAGE RETURN) = CR + erasure to end of line
CTL + H = BS (BACK SPACE) = cursor ←
CTL + [= ESC (ESCAPE) = scroll up
CTL +] = CR (carriage return) without erasure
CTL + Z = SUB = erasure of current line
↖ = FS (FILE SEPERATOR) = home cursor

CMOS FSK modulator
Summer Circuits 1978, circuit no. 72. The reset connection of the 4020 (IC1) should be connected to supply common, instead of to positive supply.

Master tone generator
September 1978, page 9-09. The M 087 is an SGS/ATES type; not Motorola, as stated in the article. Our apologies for any inconvenience this may have caused.

Databus buffer
October 1978, page 10-18. On the component layout (figure 3) the '+' and '0' supply connections are transposed: the '+' connection runs along the outer edge of the board to pins 20 of IC1 and IC2.

Tag!
November 1978, page 11-12. P3 should be a 2k2 (or 2k5) linear potentiometer.

Ring the bell and win a prize!
November 1978, page 11-22. The output voltage U_p, referred to at several points in the text, is present at the emitter of T2 (point 'M'). Furthermore, the component numbers given in the parts list are incorrect; the corrected list reads:

Resistors:
R1,R4,R19,R20,R22 = 10 k
R2,R8,R9,R10,R12,
   R17,R21 = 100 k
R3 = 33 k
R5,R6 = 2k2
R7 = 220 k
R11,R14,R15 = 1 k
R13 = 180 k
R16,R18 = 15 k
R23 = 220 Ω
P1,P3,P4 = 100 k preset
P2 = 10 k lin

Capacitors:
C1,C7,C8 = 100 n
C2,C5 = 4n7
C3 = 1μ5/16 V
C4 = 100 μ/16 V
C6,C9 = 220 n

Mini push button switches
A range of miniature snap-action push button switches, designed for printed circuit board mounting on a 0.1 inch (2.54 mm) hole grid, is now available from Impectron Ltd. Switches in the Petrick 320 range feature attractive, ergonomically designed faces and a positive snap or rocker action. Single pole press-to-break and changeover configurations are all available as standard, and an LED indicator can be incorporated to provide visual warning of function selected.
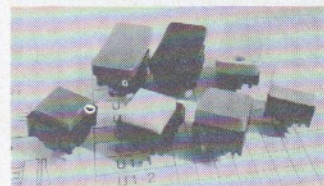
The basic body size of all types is 12.4 x 12.4 x 7.5 mm (approx. 0.5 x 0.5 x 0.3 inches) with 4 mm (0.16 in) long connection tags. The selected push button or rocker action will add a further 5 mm to body depth. This space saving design allows very compact keyboards and control panels to be constructed economically. Switches may also be mounted in line or block format to form key sets with mechanical interlocking and release mechanisms. Despite their small size the switches are capable of handling up to 25 mA at 50 V DC. Contact resistance is less than 20 milliohms and operational life is in excess of 10^5 operations at full load. Operating temperatures range is $-25°$ to $+75°$C.

Impectron Limited,
Impectron House,
23-31 King Street,
London W3 9LH

(984 M)

C-MOS 64-bit memory
RCA Solid State has introduced a new complimentary MOS 64-bit random-access memory, the CD40114B, which is equivalent to, and pin-compatible with, industry-standard TTL types such as the 7489, but offers the additional C-MOS advantages of low power consumption, high noise immunity and wide supply voltage range (20 V rating). The CD40114B is equivalent to C-MOS type 74C89.
The CD40114B is a 16-word x 4-bit random-access memory with four address inputs, four data inputs a 'write-enable' input, a 'memory-enable' input, and four 3-state data outputs. The memory-enable input and the 3-state data outputs allow memory expansion, and the four address inputs are decoded internally to select one of the 16 possible word locations.
An input address latch facility is included; the address information is latched on the negative edge of the memory-enable input by an internal address register. The selected output assumes a high-impedance condition when the device is writing or disabled.
RCA Solid State,
Sunbury-on-Thames,
Middlesex, TW16 7HW

(981 M)

DC/DC converters
Gresham Lion Limited have added a further series of miniature encapsulated DC/DC converter units to its popular GEMINI range. Designated the Gemini 800 Series, the new additions will extend the existing range by 24 units, making it one of the most comprehensive currently available.

The 800 series consists of 16 single output and eight dual output Converters operating from nominal DC inputs of 5, 12, 24 or 28 volts. Designed to operate over wide input voltage ranges, these 10 and 12 watt units provide excellent line/load regulation and high efficiency (65% typical) even with output loads as low as 10% of rated output.
All models employ Pi-type input filters to minimize reflected input ripple current. EMI/RFI protection is provided through complete six-sided continuous shielding incorporated into the design of each unit. In addition, all units in the range feature output current limiting and output short-circuit protection for up to eight hours, with automatic restart when the short circuit is removed. The units will also start with reverse polarity current injected into the output. The units are designed for PCB mounting, they measure 89 x 63 x 22 mm (3.5 x 2.5 x 0.88 inches) (L x W x H), and will operate in the temperature range -25 to $+71°$C.
Gresham Lion Ltd.,
Gresham House,
Twickenham Road,
Feltham, Middlesex.

(927 M)

MARKET

Subminiature Relay

One of the world's smallest relays, the G2E, is available with coil ratings from 1.5 V to 24 V at currents of 18.8 mA to 300 mA. This subminiature component measures only 16 x 11 x 10.5 mm and weighs just 3.5 g.
Features include long service life,
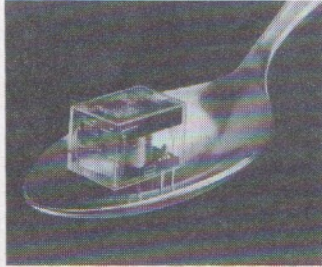
in excess of 5 x 10⁶ operations, and a switching frequency of 1800/hr max. Coil resistances range from 5 Ω to 1280 Ω and this highly sensitive relay requires only 290 mW of pull-in power. Both operate and release times are ≥ 5 ms, and the G2E has an insulation resistance of ≮ 100 MΩ (measured with a DC 500 V megger).
Designed specifically for p.c.b. use, the relay has its pins set on the international 2.5 mm grid terminal arrangement. A transparant case protects the component from dust, while the operating ambient temperature range is −25°C to +55°C.
Relay contacts are rated at 0.5 A (100 V AC PF = 1) or 1 A (24 V DC), while the carry current is 2 A for both versions.
IMO Precision Controls Limited, 349 Edgware Road, London W2 1BS

(973 M)

23 mm potentiometers with switch

The new Iskra 23 mm carbon potentiometers are spindle-operated and incorporate a switch rated at 250V AC, 4 A continuous current, 80 A surge current. Known as the P8SL series, they are available from Iskra Limited of Coulsdon, Surrey, the British trading subsidiary of the multinational Iskra organisation.
Each switch is approved to BS415 and SEMKO standards, and the series is produced in both linear and log values. The linear values extend from 100 ohms to 10 megohms, and the log values from 2.5 kilohms to 5 megohms.
Maximum voltages which can be applied to the carbon track are: linear units, 500 V DC; log units, 350 V DC.
All potentiometers in the P8

range can be supplied with any one of three tappings — 40%, 50% or 60%. Each unit has a 295-degree angle of rotation and a smooth rotary action. End-stop torque is 8 kgcm.
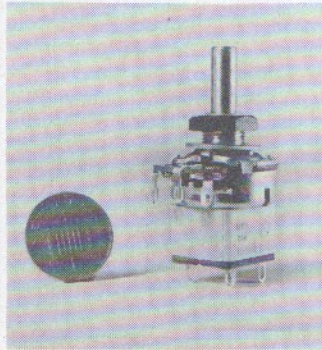
Low-noise carbon-brush wipers are standard throughout the range. Rated dissipations are 0.4 watts (linear) and 0.2 watts (log). Mechanical fixing is by a standard M10 x 0.75 mm screw thread. Two mounting styles are available, one standard and one special. The standard is designed for a 10.1 mm diameter centre hole, with a 2.1 mm (+0.1 mm, −0 mm) locating hole. On special request, Iskra can supply units for mounting in a 10 mm diameter hole with a rectangular section extending from the circumference. This section measures 2.5 mm x 2.6 mm (+0.1 mm, −0 mm).
Dual-gauged versions for stereophonic applications are also available.
Iskra Limited, Redlands, Coulsdon, Surrey, CR3 2HT

(986 M)

16.8 mm carbon potentiometers

There are four types in this new 16.8 mm spindle-operated range, and together they cover many applications.
Type P10 units are available with linear resistance values from 100 ohms to 5 megohms, and log values from 2.5 kilohms to 5 megohms. Rated dissipations
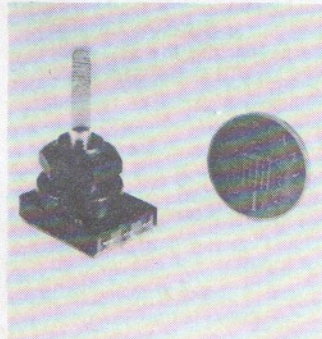
are 0.15 watts (linear) and 0.08 watts (log). Designed for mounting with a M7 x 0.75 mm screw, all P10 potentiometers incorporate a metal contact wiper. Angle of rotation is approximately 295 degrees, and all units have a smooth rotary action. End-step torque is equal to or greater than 4 kgcm.
All potentiometers in the P10 range are available with an integral switch rated at 12 V, 2 A. When a switch is incorporated, the type number is designated P10S.
For stereo hi-fi and similar applications, Iskra has produced a tandem version of the P10. Called the P10T, it has the same specification as other units in the P10 range. Matching can be achieved down to 26 dB.

16.8 mm series with plastic cases

Many applications today demand a higher resistance to the ingress of dust and other harmful particles. To cater for this demand, Iskra Limited has introduced a series of plastic-cased potentiometers. Known as the P11P series, these 16.8 mm spindle-operated carbon units are specifically designed for printed-circuit applications, but are also available with solder terminations. As well as featuring a plastic case, all P11P units have a 4 mm plastic spindle and a plastic threaded sleeve (M10 x 0.75 mm). Resistance values are 100 ohm to 5 megohms linear, and 2.5 kilohms to 5 megohms log. The angle of rotation is approximately 270 degrees. Each P11P potentiometer has a smooth rotary action, and an end-stop torque of 4 kgcm. Rated dissipations are 0.15 W linear, and 0.08 W log.
Iskra Limited, Redlands, Coulsdon, Surrey, CR3 2HT

(987 M)

P.C.B. mounted illuminated push-button switches

Highland announce the release of Series 99 illuminated push-button switches and indicators for printed-circuit-board mounting. Designed for applications on both industrial and domestic equipment, this new range has low-level switching elements rated 100 mA, 50 V max. Push buttons and indicator assemblies are square — screen size 18.6 mm square with both flat and concave illuminated screens.
The switch and indicator terminations are by 1 mm pins for direct p.c. board mounting. They may be flow soldered. Illumi-

nation is provided by miniature T1 size bi-pin lamps available in 6, 12, 18, 24, 28 and 36 Volt versions.
The range comprises Indicators (signal lamps) and both momentary and maintained-action switches with low-level switching elements having a 1 make/1 break (non-shorting) contact configuration.
Film legends, designed and produced to customer's requirements by Highland, are intended for use on Series 99 devices. These are sandwiched between the clear outer screen and the inner white diffusor with a colour backing film if required.
Highland Electronics Ltd., Highland House, 8 Old Steine, Brighton, East Sussex, BN1 1EJ
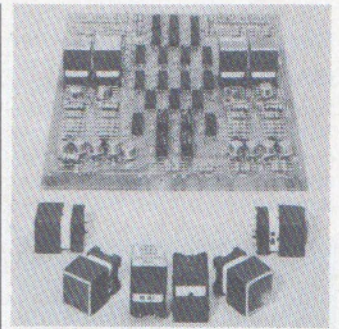
(988 M)

The plug which thinks for itself

This ingenious three pin 13 amp plug by means of neons and indicator on the front of the plug shows immediately if:
1. Socket is A1 OK
2. DANGER reverse polarity
3. DANGER no earth
4. DANGER live fault
5. DANGER neutral fault
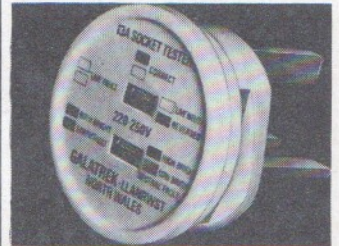
This tester is universally needed by DIY buffs, craftsmen, tradesmen, householders, installers of microwave and solid state circuitry and throughout industry and commerce. It was initially designed to protect electronic machinery installed by C. & A. Modes in all their stores throughout the U.K.
Galatrek Engineering, Scotland Street, Llanrwst, Gwynedd, LL26 0AL, North Wales, Great Britain.

(975 M)

MARKET

Rectifiers for X-Ray power supplies

Semtech have introduced the 'X-Way Stic', a new series of open rectifier sticks specifically designed for X-Ray power supplies.

Each X-Way utilizes hermetically sealed 'Metoxilite' multi-chip 'avalanche' rectifiers mounted on a PCB. These 'Metoxilite' multi-chip rectifiers (initially developed for high reliability aerospace programmes) are now available at commercial prices. In addition to X-Ray power supplies, these rectifiers can be efficiently used in most standard, single and polyphase circuits.
Type: X 100KS, X 125KS, X 150KS.
PIV : 100, 125, and 150 KV.
Average Rectified Current (at 55°C oil): 220 mA.
Size : (overall) 6.22"L x
          .690"W x (end cap)
          .500"D (anode) .390"L.

Bourns (Trimpot) Limited,
Hodford House,
17/27 High Street,
Hounslow,
Middlesex.

(926 M)

Desk console

Recently introduced by BOSS Industrial Mouldings Limited is the new BIM 8005 Small Desk Console, the main feature of which is its three piece construction which allows rapid and easy removal of both top and bottom panels.
The main body of the console is manufactured in ABS, as is also the removable bottom panel which incorporates 9 mounting bosses for supporting either a full 100 x 160 mm Eurocard or 2 half size boards.
The flush mounted 1 mm thick
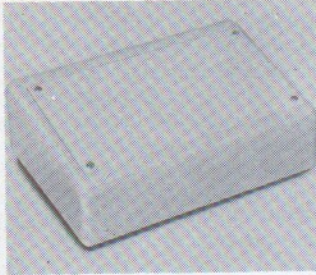
grey aluminium top panel tones with the orange, black or blue console body, the latter — as with the majority of BOSS enclosures — also incorporating slots for vertically mounting 1.5 mm thick pcb's as well as the recently announced BIMDAPTORS which allow pcb's to be mounted sandwich fashion.
Measuring internally 165 x 122 x 60 mm high and incorporating small, self adhesive, rubber stand-off feet on the base, this new small desk BIMCONSOLE is equally suited to initial development, pre-production and OEM type applications.
BOSS Industrial Mouldings Ltd,
Higgs Industrial Estate,
2 Herne Hill Road,
London, SE24 0AU, England

(921 M)

Voltage-to-frequency converters

A new family of Voltage to Frequency (V-To-F) converters, designated LM131/231/331 is now available from National Semiconductor Corporation, with better specifications than any previous device.
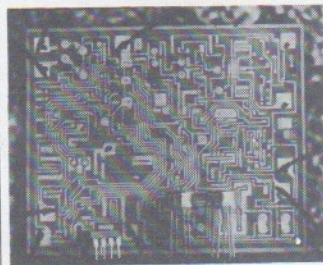
It is claimed that the new devices all function with guaranteed 0,01% maximum non-linearity as 10 KHz V-to-F or F-to-V converters, and can be operated

from a single +4.0 to +40 volt power supply.
The standard device is available with a gain temperature coefficient (T.C.) of 150 parts per million/°C typical (150 ppm/°C maximum). These devices are particularly suited to high-resolution data acquisition applications. The premium-grade versions of the LM131 family (LM131A/231A/331A) also feature a guaranteed gain T.C. of ± 50 ppm/°C maximum.
The LM 131A is the first voltage to frequency converter in the industry guaranteed to have a 50 ppm/°C T.C. spec over the entire −55°C to +125°C temperature range.
Designed to function as an improved, pin for pin replacement for the 4151 V-to-F converter, the LM131 is specified to operate over its 4 to 40 volt supply range with improved accuracy.
The LM131 also features low power dissipation (14 milliwatts) when operated from +5 volt supplies, and this is highly suitable for battery powered applications.

National Semiconductor,
301 Harpur Centre, Horne Lane,
Bedford MK40 1TR.

(909 M)

Temperature range	
−55° to 125°C	LM131 H - LM131 AH
−25° to +85°C	LM231 H - LM231 AH
0° to +70°C	LM331N and
	LM331 H - LM331 AH

High-voltage video transistors

Micro Electronics Ltd. has introduced a wide range of high-voltage n-p-n silicon planar transistors designed for video-amplifier applications in colour and monochrome television receivers. Available in TO-39 and TO-92 cases, the devices feature collector-emitter breakdown voltages up to 300 V and have good frequency characteristics, with current-gain bandwidth products of 50 MHz. Devices in the TO-39 package include the BF 257/258/259, designed for high-voltage video output stages, and the BF 336/337/338, designed for R-G-B and coulour-difference output circuits. TO-92 packaged devices include the BF 297/298/299 and BF 391/392/393 transistors.
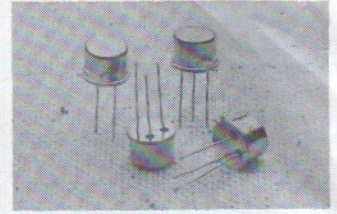
The devices are designed for maximum emitter-base voltage of 5 - 8 V, and cover maximum collector currents from 100 mA to 500 mA. Total maximum power dissipation at a case temperature of 25°C ranges from 1.5 W to 5 W.
Micro Electronics Limited,
York House,
Empire Way,
Wembley, Middlesex.

(918 M)

Low-noise audio transistors

A new range of small-signal transistors from Micro Electronics Ltd. are designed for low-noise preamplifier, audio-frequency amplifier driver stages and signal-processing circuits in television receivers. Designated the BC 413/414/415/416 Series, the transistors are low-cost silicon planar epitaxial devices featuring high breakdown voltage, high gain and low flicker noise.
The BC 413 and BC 414 are n-p-n transistors, while the BC 415 and BC 416 are the complementary p-n-p devices.
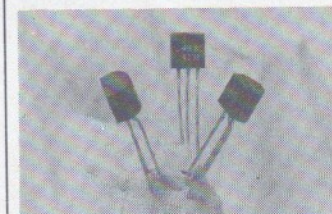
Maximum collector-base voltage is 45-50 V, maximum emitter-base voltage 5 V. Maximum continuous collector current is 100 mA, and total power dissipation is 300 mW.
Current gains ranging from 125 up to 900 are available, typical noise figure is 1.2 dB, and flicker-noise voltage referred to base is only 0.11 μV for the BC 415 and BC 416 and 0.135 μV for the BC 413 and BC 415.
Micro Electronics Limited,
York House,
Empire Way,
Wembley, Middlesex.

(930 M)

MARKET

µP appliance timer

A single package MOS-LSI microcircuit appliance timer, consisting of a central processor with on-chip memory, has been introduced by General Instrument Microelectronics. The circuit – basically a 4-bit microprocessor – is available as a basic 28-lead or a more sophisticated 40-lead version, and will considerably widen the scope of domestic appliance designers. The device is essentially a versatile, low cost timer, providing designers with the type of facilities necessary for controlling cookers, driers, central heating, etc. The 28-lead version designated AY-3-1250, accepts instructions from 'hours up', 'hours down', 'minutes up', or 'minutes down' keys, where momentary depression of keys cause single increments or decrements, and continuous depression causes the displayed digits to cycle. In use the circuit is linked to a 4-digit LED display indicating any function selected. It has 3 separate outputs for which on and off times may be programmed in.

In a cooker application, all normal instructions may be programmed in, including ON time, OFF time, cook duration, etc. Other facilities include a 'minute minder' with audible alarm output, indication of mains failure, 12/24 hour operation, temperature setting or an optional fluorescent display driver output interface.
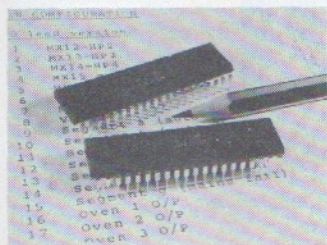
The 40-lead version – designated AY-3-1251 – is designed for more sophisticated systems where 10 x 4 keyboard or touchpad entry and 14-digit permanent display are required. It has four controlled outputs, each with a variable mark-space ratio for control of hotplate duty cycles, etc. The 14-digit display facility could be used for a minute minder (3 digits), oven temperatures (3 digits), time on/off (4 digits), and hotplate temperature (4 digits).

When used for fully automatic cooker control, the time programme would be entered and cooking temperature selected using a key pad. When the start time is reached the appropriate output would be activated and an 'ON' indicator lamp energised.

When the stop time is reached the output is deactivated and the minute minder audible alarm activated for 10 seconds. All 3 programmable outputs may be separately controlled in this way, but the device can also be used in a semi-automatic or manual mode. A further facility allows a set programme to be repeated at 24 hour intervals by the simple depression of a 'repeat' key.

Both 28-lead and 40-lead versions include a built-in standby frequency source, which allows the devices to function normally during mains failure. In this event the circuit detects the absence of 50/60 Hz input, and a 200 kHz oscillator takes over timing under external battery power and lights a 'mains failure' warning lamp. On the re-application of the mains supply the circuit returns to mains power, but leaves the lamp on (until it is manually reset) to indicate that mains failure has occured.

Although the two versions described here are standard devices, the microcircuit chip is mask programmable to include a wide range of options and output requirements. These include steady or pulsed temperature display outputs, error 'E' display, programmable time delays, touchpad inputs. Both versions are extremely suitable for general control applications requiring simple arithmetic, display of up to 43 digits and interfacing with analogue systems.

General Instrument Microelectronics Ltd.,
Regency House,
1-4 Warwick Street,
London W1R 5WB, England

(982 M)

Video monitoring oscilloscope

Gould Instruments Division has won a substantial order from the British Broadcasting Corporation for the supply of an oscilloscope designed for video monitoring applications. The oscilloscope is a modified brighter version of the Gould Advance OS3300B with a BBC designed timebase module incorporating comprehensive video triggering facilities, which is being made by Gould under a manufacturing licence agreement from the BBC.

The new timebase generator allows the oscilloscope to be used for detailed line-by-line examination of 625-line television waveforms or to display a television picture. It accepts a standard level video signal, which may contain 'Sound-in-Sync'

signals and provides six different triggering modes: field 1; field 2; field 1 and 2 alternating; line repetitive; single line selectable by front panel switches (with the line number indicated on a 3 digit light emitting diode display) and line pairs in the range 16/329 to 22/335.

The triggering can be delayed continuously by up to 90 μs via a multiturn potentiometer, which allows the signal to be examined in detail. The displayed video signal may be clamped or not, as required. When the unit is used to display a television picture, the triggering point selected may be observed as a 'bright up line' on the picture, enabling the waveforms to be rapidly related to the picture. The changeover from waveform to picture is effected by a single front panel switch. The modified timebase retains its normal triggering facilities, so that the instrument may be used as a general purpose single timebase oscilloscope.

Gould Instruments Division, Roebuck Road, Hainault, Essex

(974 M)

1 MHz to 1600 MHz in one sweep

Gould Instruments Division has entered the sweep-generator market with the introduction of the Gould Advance SW100, a high-performance instrument which can cover the frequency range from 1 MHz to 1600 MHz in a single, flat sweep. Other key features of the instrument include a comprehensive automatic marker system for precise frequency determination and an output level unit that can be varied in 0.01 dB steps.

The Gould Advance SW100 incorporates electronic switching between the three basic ranges (1-500 MHz, 450-1000 MHz and 950-1600 MHz) to give the effect of a single sweep. Deviation from flatness is ± 0.25 dB using a power meter, or ± 0.35 dB measured on a full-wave detector. The three operating modes of the SW100 are: start/stop or F1/F2 mode, in which a sweep occurs

from one set frequency to another; FC/ f (delta F) mode, in which a centre frequency is set in the middle of a spectrum of interest with one control and a deviation about the centre frequency is set with another control; and signal generator mode, in which the sweep drive is switched off and the sweeper is used as a straight signal generator. An extremely accurate output attenuation system offers ten steps of 10 dB, 1 dB, 0.1 dB or 0.01 dB, and the output level range can be varied from +6.66 dBm (471 mV or 4.6 mW) to −103.33 dBm (3 μV or 0.18 pW) in 0.01 dB steps.
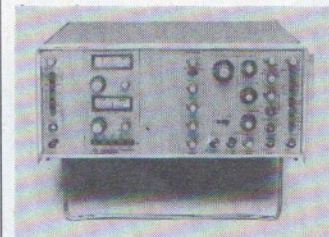
The SW100 incorporates a unique high-accuracy automatic harmonic marker system with frequencies of progressively increasing resolution being cut in as they are needed. Markers are provided with 100 MHz, 1 MHz and 100 kHz separations, and the markers can be oriented in horizontal, vertical or 45° directions to prevent them being obscured by the waveform under examination. A unidirectional marker facility is also provided for cases where the instrument is used with low-speed output devices such as X-Y plotters. Internal and external amplitude and frequency modulation (both fully swept) are available, and comprehensive internal and external sweep-rate controls are incorporated. The sweep may be varied between 70 Hz (14 ms sweep) and 120 seconds per sweep. A manual sweep control is also incorporated.

Harmonic-related spurious outputs are typically −35 dB, and nonharmonic spurious outputs are less than −40 dB. Linearity is better than ± 2%.

Construction of the SW100 is based on a modular approach, with easy accessibility for servicing and maintenance, leading to low cost of ownership. Three blank panels are provided on the front panel to accommodate future options; a hard-copy output option is already available. The Gould Advance SW100 measures 320 x 145 x 365 mm, and weighs 8.6 kg.

Gould Instruments Division, Roebuck Road, Hainault, Essex

(980 M)